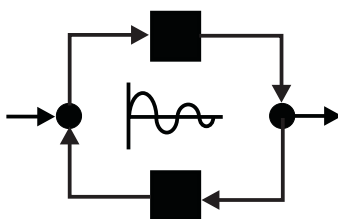




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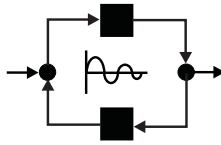
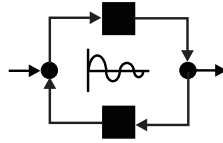


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A Bi-Directional DC- DC Converter with Adaptive Fuzzy Logic Controller

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Abstract:

This paper introduces a Bi-directional DC-DC converter with adaptive fuzzy logic controller (AFLC). Bidirectional power flow is obtained by same power components and provides a simple, efficient, and galvanically isolated converter. In the presence of DC mains the converter operates as buck converter. Powers the down stream converter and also charges the battery. When the DC mains fails, the converter operates as boost converter and the down stream converter is fed by the battery. In both the modes the power switches are controlled by Pulse Width Modulation (PWM) technique and the pulses are generated by the application of fuzzy logic with an adoption algorithm. The proposed converter is simulated using MATLAB and laboratory prototype is developed to validate the simulation results

Keywords: Bi-Directional Power flow, Fuzzy logic control, DC-DC converter

1. Introduction:

The Bi-directional DC-DC converters are involved in power flow between two dc sources. They allow power flow in both the direction without change in polarity of voltage. They are used in DC Un-interruptible Power Supplies and Battery charging circuits. In literature, bi-directional power flow is obtained by constructing individual converters for each direction of power flow [5]. Then they are implemented by soft switching [7] and resonant techniques [6]. But it increases the component count, circuit complexity also its losses. It misses the soft switching signals at light loads.

To overcome these difficulties a Bi directional converter which is a merge of a half bridge converter and push pull converter is proposed. It claims the following advantages. They are (i) Low stress on switches ii) Galvanic isolation (iii) Reduced components count. MOSFET switch is used in this converter. The bidirectional power flow property of MOSFET enables power flow of the proposed converter in either direction using only one transformer. The dynamic behavior of this converter is also improved by the fuzzy logic.

Generally any of the intelligent technique improves the dynamic performance of the converter. The Neural Networks (NN) can be seen as an alternative used to

model and control nonlinear and linear systems where the traditional methods fail [2]. The problem in neural networks design is often seen on two perspectives: the first is the number of examples necessary to obtain a good generalization; the second is the size of the neural network. The identification of the parameters of the NN, otherwise called training, is often carried out by the back propagation algorithm [4], which is based on the minimization of the error of training and the chaining rule. This algorithm uses a gradient decent showed several disadvantages such as the slowness of convergence, the sensitivity to the local minima and the difficulty in regulating the training parameters. approaches were proposed to improve the back propagation algorithm: modification or decentralization of the step size, use of quasi-Newton algorithms [4], algorithms genetic [4] etc. On the other hand, the approximation capability of a neural networks is closely related the number of the neurons in the hidden layer. One usually seeks a compromise between the accuracy and the complexity of the network. Therefore fuzzy logic is chosen for converter control in this paper.

The fuzzy based controller designing strategy started nearly a decade back. In 1994 a FLC controller [10] is proposed for simple DC-DC converters. This has been implemented by using Digital Signal Processor (TMS320C50) in 1996[9]. In the year 1997 neural network control is designed for DC-DC converters[4]. In 2006 different control strategies have been proposed to control DC-DC converters and implemented by using microcontroller and a specialized hardware. In recent years neurofuzzy controllers and adaptive fuzzy logic controllers simplify the process of controller design and provide efficient control. More interest has been developed in the intelligent technique based controllers for most of the circuits because of i) simplicity in controller development ii) Possibility of automated control iii) Need of less skilled labour.

This paper proposes a new topology of DC/DC converter with bidirectional power flow which reduces the cost of the converter and improves the efficiency. The intelligent technique based controlling strategy improves the dynamic behaviour of the proposed converter.

The paper is organized as follows: Section 2 presents the description and operating modes of the proposed converter. Design considerations of the controller are



discussed in Section 3. Section 4 explains the simulation results and various current and voltage waveforms obtained by simulation. Section 5 represents the experimental results of the proposed converter constructed

2. Proposed converter topology

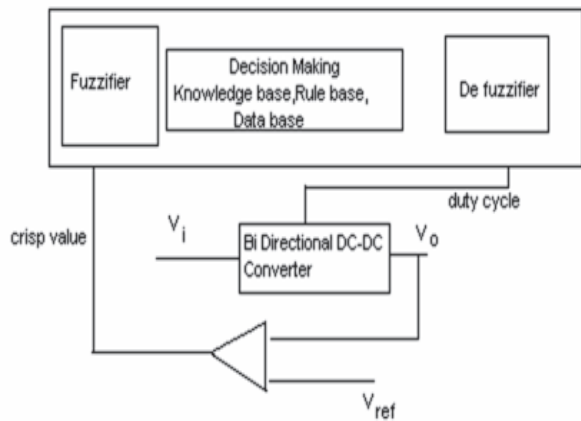


Figure1. Block diagram of Fuzzy logic L Controlled Converter.

This paper proposes a bi-directional DC-DC converter controlled by Adaptive Fuzzy Logic Controller. Adoptive fuzzy logic denotes fuzzy logic with an adoption algorithm. Fuzzy Logic controller is constructed and adoption algorithm is imposed on this controller. The overall block diagram of the proposed converter is shown in figure.1. This includes the internal blocks of the fuzzy logic controller, bidirectional DC/DC converter with sensing and error producing component.

The block diagram of bidirectional converter proposed in this paper is shown in figure 2. It includes two converters topologies for power flow in either direction. Both the topologies are involved in power flow to the load irrespective of the presence of mains supply. The power circuit of the converter is shown in figure 3.

In the circuit proposed, primary side of the transformer is connected with a half bridge converter which includes the switches S1 and S2 and fed by power supply. The secondary of transformer is connected to the switches S3 and S4 and forms a push pull topology. The converter operates in two modes. In forward mode operation switches S1, S2 are operated and S3, S4 are switched off. The converter exhibits the operation as buck converter. In back up mode S3, S4 are switched and the switches S1, S2 are in off condition. Operation of the converter in this mode is similar to boost converter. Individual switching signals applied for individual switches

Following assumptions were also made for proper analysis of the converter.

- The switches and diodes are ideal
- Transformer is ideal and the turns ratio is one
- All the voltages and currents applied are periodic

The proposed half bridge and pushpull combined topology is advantageous in the following aspects when compared to the conventional converters. (i) The switches are subjected to voltage stress only equal to the DC input voltage. But in the conventional push pull converter the

switches are subjected to voltage stress twice the input

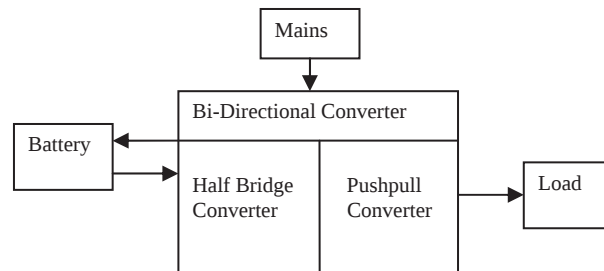


Figure 2. Block diagram of proposed converter

voltage. (ii) In low power conditions usage of two switch half bridge topology reduces the size of LC filter required, when compared with four switch full bridge converter topology.

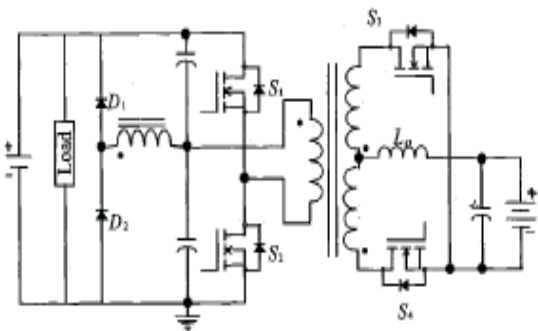


Figure 3. Power Circuit of Proposed Converter

As per the block diagram of the converter shown in figure.2, the switches are controlled by the error signal produced. So the required duty cycle variation of the switching signal depends on the input voltage of the converter and battery voltage. It is given by the expressions shown below.

Forward mode

$$d_{\min} = \frac{V_{\text{battmax}}}{V_{\text{smin}}} \quad d_{\max} = \frac{V_{\text{battmax}}}{V_{\text{smax}}}$$

Back up mode

$$d_{\min} = \frac{V_s - V_{\text{battmin}}}{V_s} \quad d_{\max} = \frac{1 - 2V_{\text{smin}}}{V_s}$$

$d_{\min}$ -minimum value of duty cycle; $d_{\max}$ -maximum value of duty cycle, V_{battmax} -Maximum value of battery voltage, V_{battmin} -Minimum value of battery voltage V_{smin} -Minimum value of mains supply voltage, V_{smax} -Maximum value of mains supply voltage, V_s -Output voltage in backup mode.

3. Controller design

The process of fuzzy logic controller design includes the following steps. (i) Fuzzification; Process of representing the inputs as suitable linguistic variable. (ii) Decision Making; Appropriate control action to be carried out. It is based on the knowledge base and rule base. Knowledge base and rule base are the details about the linguistic variables and control rules (iii) Defuzzification; Process of converting fuzzified output into crisp value. The inputs to the FLC are error signal and difference of error signal. The output is the duty ratio of the switching signal.



$$e = V_{ref} - V_o; \quad d(t) = d(t-1) - d(x(t))$$

$d(t-1)$ =Duty cycle at $(t-1)^{th}$ instant; V_o =Output Voltage;
 $d(x(t))$ =Change in duty cycle; V_{ref} =Reference Voltage;
 $d(t)$ =Duty cycle at t^{th} instant; e =error signal;

All the linguistic variables are assumed to have same number of linguistic values. The Shrinking span membership function algorithm proposed by Chies and Hsieh is used to construct FLC without need of expert presence. This algorithm involves the process of arranging the membership functions of a linguistic variable in orderly manner across universe of discourse. The shrinking factor (S_f) decides the span of membership function. By applying various values of S_f to one linguistic variable, most suitable membership function can be decided.

Mamdani type controller is chosen for this application and the basic rule of this type of controller is

IF e is A and de is B THEN $d(t)$ is C

Here A and B are fuzzy subsets and C is fuzzy singleton. Each universe of discourse is divided into seven subsets such as Positive Large (PL), Positive Medium (PM), Positive Small (PS), Zero (ZE), Negative Small (NS), Negative Medium (NM), Negative Large (NL).

Table 1. Rule mapping by index representation

e/de	-3	-2	-1	0	1	2	3
3	0	1	2	3	4	5	6
2	-1	0	1	2	3	4	5
1	-2	-1	0	1	2	3	4
0	-3	-2	-1	0	1	2	3
-1	-4	-3	-2	-1	0	1	2
-2	-5	-4	-3	-2	-1	0	1
-3	-6	-5	-4	-3	-2	-1	0

Table 2 Labels of linguistic variables

e/de	NB	NM	NS	ZE	PS	PM	PB
PB	ZE	PS	PM	PB	PB	PB	PB
PM	NS	ZE	PS	PM	PB	PB	PB
PS	NM	NS	ZE	PS	PM	PB	PB
ZE	NB	NM	NS	ZE	PS	PM	PB
NS	NB	NB	NM	NS	ZE	PS	PM
NM	NB	NB	NB	NM	NS	ZE	PS
NB	NB	NB	NB	NB	NM	NS	ZE

The rule base for the corresponding membership functions is decided by index representation method as shown in table 1. Table 2 shows the linguistic label representation of rulebase.

The center of gravity method of defuzzification is applied to obtain crisp result from the linguistic values obtained according to rule base.

$$du = \frac{\sum W_i C_i}{\sum W_i}$$

du denotes the change in duty ratio inferred by the i^{th} rule. C_i denotes fuzzy singleton and W_i denotes weighting factor. The inputs to the AFLC are the training

data from model file and the outputs are the membership functions. It adopts various parameters according to the training data fed through pattern file. The duty ratios of all four switches must be determined in order to calculate other circuit parameters. The constraints to be borne in mind are that the output (battery) voltage in the forward mode can at most be equal to half the input dc bus voltage and in the backup mode the output voltage (at the dc bus) must be atleast equal to the input (battery) voltage. Also, the maximum theoretical duty ratio of S_1 and S_2 in the forward mode must be less than 0.5 while the minimum theoretical duty ratio of S_3 and S_4 in the backup mode must be greater than 0.5. The duty ratios are dependant on the voltages at the input and the output in both operating modes. The algorithm calculates the overall error index (e_{io}) and updates the parameters to make e_{io} to be zero. Finally application of AFLC ensures desired operation of proposed converter.

4. Simulation Results

The proposed converter is simulated using MATLAB package. The rating of converter in both the modes given in Table 3

Table 3 Rating of converter in both the modes.

Parameter/Mode	Forward Mode	Backup mode
Input Voltage	300-400V	No mains supply
Output voltage	300-400V	350V
Output Power	100W	300W
Operating Frequency	100KHz	100KHz

The waveforms are repetitive for every switching cycle. The waveforms is divided into many time intervals and analyzed.

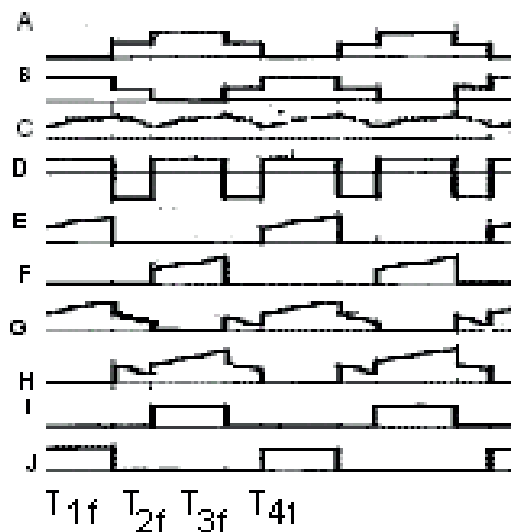


Figure 4. Waveforms during forward mode operation
A) Voltage across switch S_1 B) Voltage across switch S_2 C) Load current D) Load Voltage E) Current through S_1 F) Current through S_2 G) Current through body diode of S_3 H) Current through body diode of S_4 I) Voltage across switch S_4 J) Voltage across switch S_3



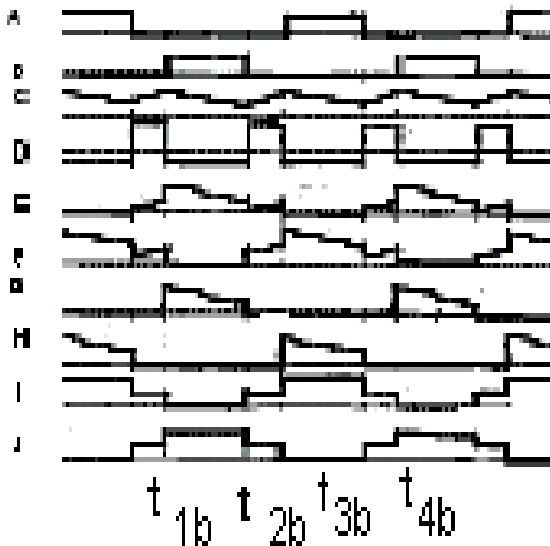


Figure 5. Waveforms during backup mode operation

A) Voltage across switch S_3 B) Voltage across switch S_4 C) Load current D) Load Voltage E) Current through S_3 F) Current through S_4 G) Current through body diode of S_1 H) Current through body diode of S_2 I) Voltage across switch S_2 J) Voltage across switch S_3

Forward mode operation is divided into four time intervals. The waveforms are shown in figure 4. During interval t_{1f} $V_s/2$ voltage appears across primary winding and the primary current builds up. In interval t_{2f} there is no voltage across primary and secondary winding. So there is no power transfer and the converter performs freewheeling action. Half the input voltage appears across switches. In the third interval t_{3f} the operation is similar to interval t_{1f} and the secondary side diodes offers rectification. In the interval t_{4f} the operation is similar to interval t_{2f} . There is no primary side conduction in this interval. In all these time intervals the reverse voltage appears across switches does not exceed $V_s/2$.

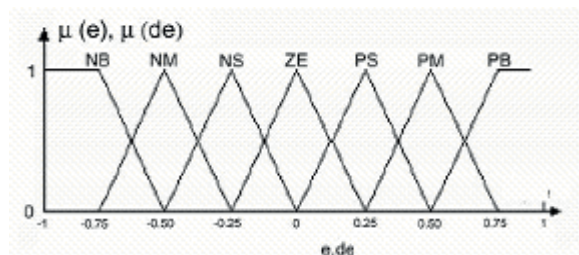


Figure 6(a) Input vectors; Error and change in error

Back up mode operation is also divided into four intervals. The waveforms are shown in figure 5. In the first interval t_{1b} , transformer secondary subjected to short circuit. The inductor stores energy and the total battery voltage appear across inductor. The bulk capacitors provide output load power. In the second interval t_{2b} the energy stored in the inductor is transferred to load. Equal voltage appears across primary and auxiliary winding. So both the capacitors get charged simultaneously. In the interval t_{3b} the operation is similar to interval t_{1b} . In the last time interval t_{4b} operation is similar to t_{2b} interval. The diodes causes equal charging of capacitors. In the simulation of AFL controller, pattern file is generated initially. It consists of three vectors two inputs

error and change of error and the output vector is change of duty cycle. Seventy two epochs were performed and the error measure is 9.3×10^{-5} . The shrinking factor sf value is 0.35. Figure 6(a)&6(b) show the waveforms of input and output vectors obtained during simulation.

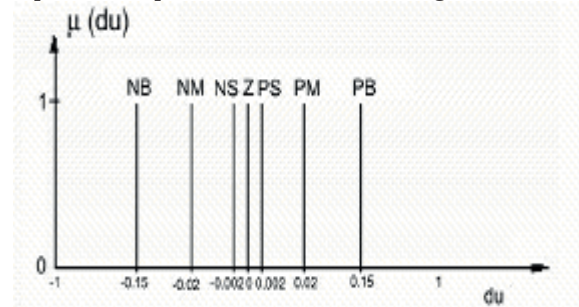


Figure 6 (b) Output vector i.e change in duty cycle

The results indicate that this method provides an easy and systematic way in designing the AFLC. The generated membership functions and rule base are general and could be used for any without any modification.

5. Hardware results

Laboratory prototype of the above proposed converter is developed with the help of microcontroller ST52T420. It is an 8 bit micro controller. It effectively performs fuzzy logic and Boolean operations. Components used in prototype construction are tabulated in Table 4. The static performance was studied for both the forward and backup modes. The dynamic response of the converter under transient conditions of step changes in load and Switchover from charging to backup mode is illustrated.

Table 4- Specifications of Components used in Prototype.

S. No	Component	Rating/Model
1	Diode	IN4007A
2.	MOSFETS	IRF840
3.	Capacitors	150 μ F
4.	By pass capacitor	470 μ F
5.	Inductors	2mH&380 μ H

Wave forms of various currents and voltages are obtained for both the modes of operation. Figure 7 shows the photograph of the converter constructed.



Figure 7. Topview photograph of the converter constructed



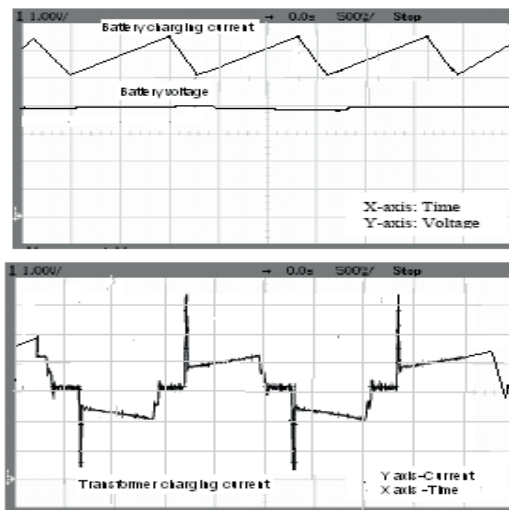


Figure 8. Battery charging current battery voltage and transformer charging current waveforms in forward mode operation

Figure 8 shows the waveforms of battery voltage, inductor current and transformer charging current in the forward mode operation when there is power available in mains. Figure 9 shows the waveforms in back up mode operation Figure 10 shows the waveforms of converter when its switching from forward to back up mode.

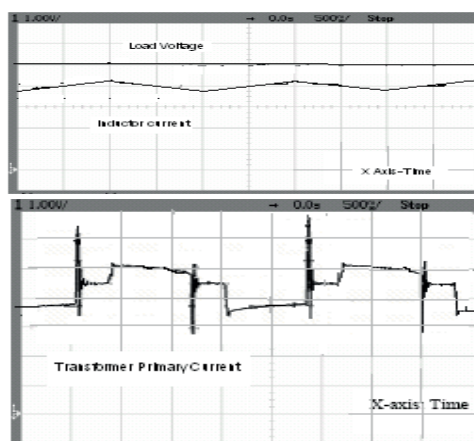


Figure 9. Back up mode Load voltage, inductor current, transformer primary current

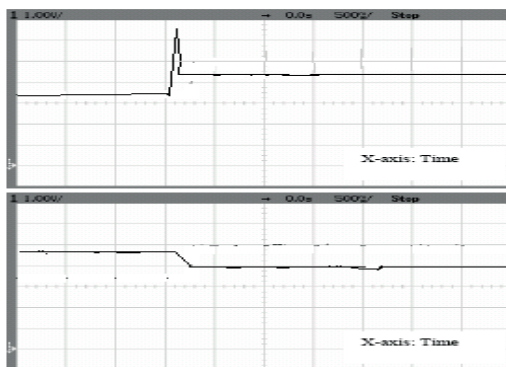


Figure 10. Waveforms of inductor current and load voltage when switching from forward to back up mode of conduction.

Figure 8 shows the experimental waveforms for the forward/charging mode, at 75% load, with the converter operating at 100 kHz (360 V, battery power 85W). The dc mains supplies load power and charges the battery to 54.5V, the battery charging current 1.55A is constant with relatively small ripple as desired.

Figure 9 shows the relevant waveforms during steady state operation of the converter in the backup/current-fed mode at 75% rated load. The battery voltage, 50V and load power, 190W. The battery discharges to boost the voltage level of the dc bus to 325V, thereby powering the load. Current spikes are observed in the transformer primary current, in the forward mode and backup modes. These are due to the reverse recovery of the diodes. These spikes are not present in the simulated waveforms because the diodes considered in simulation are ideal diodes. The higher voltage drop across them will not greatly effect the converter efficiency because of the low currents flowing through them in both operating modes when providing load rectification. Certain applications may require the converter to start operating in the backup mode when the hold up capacitors are not charged. Under such conditions, as the duty cycle of and is increased to build up the load (dc bus) voltage, current in the inductor continues to rise with the switches operating at maximum duty cycle. This continuous increase in the current results in a switch current that exceeds the rated value and can permanently damage the switch. This problem can be avoided by adding a parallel combination of a relay and resistor, in series with the battery when starting up in the backup mode, with no output voltage. The series resistor limits the previously increasing inductor current. Once the output capacitors are charged the resistor is bypassed by the relay.

Transient performance of the converter is evaluated under the condition of Switchover from Forward to Backup Mode When the Battery Is Charged: Figure 10 shows the switchover at 75% load, with the battery charged at 53 V and therefore drawing minimal current. The bus voltage is 360 V in the forward mode and is regulated at 325 V in the backup mode, which is within the load converter's working input voltage range of 300–400 V. On ac mains failure, the voltage at the dc bus starts to drop. As soon as the bus voltage is detected below 325 V, the converter begins operation in backup mode and regulates the bus voltage at 325 V. Other values of the regulated bus voltage in the backup mode are possible by appropriate changes in the line and bulk detection circuitry. The figure 10 shows that inductor current, changes direction and reaches a steady state value in a very short duration. This is due to the non ideal components involved in prototype construction. But in simulation circuit this shows the smooth variation and the converter thus provides seamless transition from forward to backup mode.

6. Conclusion:

In this paper the design methodology of a bidirectional DC-DC converter with adaptive fuzzy logic controller is proposed and evaluated by constructing the laboratory prototype. The adaptive controller is implemented on a 8



bit microcontroller. Experiments results of the converter shows the effectiveness of described adaptive fuzzy logic controller and satisfactory results without pre-constructed rules of any expert. The prototype shows good steady state operation and smooth switchover between modes. It shows low part count and galvanic isolation due to its bidirectional powerflow mode.

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Optimization of Power System Stabilizers Relying on Particle Swarm Optimizers

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Abstract

A classical lead-lag power system stabilizer is used for demonstration in this paper. Initially single first-order phase compensation block is considered. The stabilizer parameters are selected in such a manner to damp the rotor oscillations. The problem of selecting the stabilizer parameters is converted to a simple optimization problem with an eigen value based objective function and it is proposed to employ simulated annealing and particle swarm optimization for solving optimization problem. The objective function allows the selection of the stabilizer parameters to optimally place the closed-loop eigen values in the left hand side of the complex s-plane. The effectiveness of the stabilizer tuned using the best technique, in enhancing the stability of power system. The performance of the system is analyzed by using both the techniques. Stability is confirmed through eigen value analysis and simulation results and suitable heuristic technique will be selected for the best performance of the system.¹

Keywords: Rotor oscillations, Power system stability, Robust control, Simulated Annealing, Particle swarm optimization.

1. Introduction

During changes in operating conditions, oscillations of small magnitude and low frequency often persist for long period of time and in some cases even present limitations on power transfer capability. Power system stabilizer (PSS) is designed to damp the low frequency oscillations of power system [1].

PSS is used to add damping to the generator rotor oscillations by controlling its excitation using auxiliary stabilizing signals. The widely used conventional power system stabilizer (CPSS) is designed using the theory of

phase compensation and introduced as a lead-lag compensator [2].

An inter connected power system, depending on its size, has hundreds to thousands of modes of oscillation. In the analysis and control of system stability, two distinct types of system oscillations are usually recognized. One type is associated with units at a generating station swinging with respect to the rest of the power system. Such oscillations are referred to as local plant mode oscillations. The frequencies of these oscillations are typically in the range of 0.8 to 2.0 Hz. The second type of oscillations is associated with the swinging of many machines in one part of the system against machines in the other parts. These are referred to as inter area mode oscillations, and have frequencies in the range of 0.1 to 0.7 Hz. The basic function of PSS is to add damping to both types of system oscillations. Other modes which may be influenced by PSS include torsional modes and control modes such as the exciter mode associated with the excitation system and the field circuit [3]. The over all excitation control system is designed so as to maximize the damping of the local plant mode as well as inter-area mode oscillations with out compromising the stability of other modes and to enhance system transient stability.

Input to PSS is rotor speed deviation which results damping torque. In this paper the state space model of the system with PSS and AVR is designed. The time constants in the PSS are tuned using simulated annealing and particle swarm optimization. From the simulation results, the optimum design of PSS is obtained by using the best technique.

2. Controller Design

Damping torque is produced to overcome rotor oscillation. The action of a PSS is to extend the angular stability limits of a power system by providing supplemental damping to the oscillation of synchronous machine rotors through the generator excitation [4].

¹ This study has been implemented on MatLab 7.0 platform at Power System Simulation Lab, Anna University (GCT Campus)



Controller is designed to compensate lag between exciter input and electrical torque.

$$\frac{\Delta T_{PSS}}{\Delta V_S} = K \angle -\theta$$

The amount of damping introduced depends on the gain of PSS transfer function at that particular frequency of oscillation.

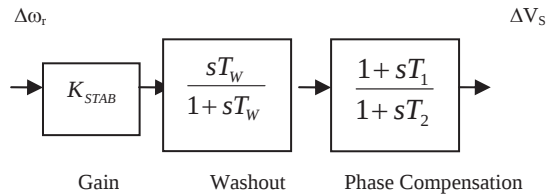


Figure 1. Lead – Lag Power System Stabilizer

Phase Lead Compensation:

To provide damping, the stabilizer must produce a component of electrical torque which is in phase with speed variations. Therefore, the PSS transfer function should have an appropriate phase-lead characteristics to compensate for the phase lag between the exciter input and the electrical torque.

The phase characteristic to be compensated, changes with system conditions. Therefore, compromise must be made and a characteristic acceptable for a desired range of frequencies (normally 0.1 to 2.0 Hz) and for different system conditions is selected. This may result in less than optimum damping at any one frequency. Generally, slight under compensation is preferable to overcompensation so that both damping and synchronizing torque components are increased.

Stabilizing Signal Washout:

The signal washout function is a high pass filter which removes dc signals, and without it steady changes in speed would modify the terminal voltage. The washout time constant is in the range of 1 to 20 seconds. For local mode oscillation, a wash out of 1 to 2 sec is satisfactory. From the view point of low frequency inter area oscillations a washout time constant of 10 sec or higher may be required in order to reduce phase lead at low frequencies.

Stabilizer Gain:

The stabilizer gain (K_{STAB}) is chosen by examining the effect for a wide range of values. Ideally, the stabilizer gain should be set at a value corresponding to maximum damping. Gain is set to a value which results in satisfactory damping of the critical system mode(s) without compromising the stability of other modes, or transient stability, and which does not cause excessive amplification of stabilizer input signal noise.

Stabilizer Output Limits:

In order to restrict the level of generator terminal voltage fluctuation during transient conditions, limits are imposed on the PSS output. The effect of the two limits is to allow maximum forcing capability while maintaining the terminal voltage within the desired limits [4].

3. Problem Formulation

In this section, the eigen value-based objective function used to robustly select the PSS parameters [5], and the

optimization problem solved with simulated annealing and particle swarm optimization.

Consider the problem of determining the parameters of a stabilizer that relatively stabilizes a family of N plants,

$$\dot{X}(t) = A_k X(t) + B_k U(t); \quad k = 1, 2, 3, \dots, N \quad (1)$$

Where, $X(t) \in R^n$ is the state vector and $U(t) \in R^m$ is the control vector.

Very often, the closed-loop modes are specified to have some degree of relative stability. In this case closed-loop eigen values are constrained to lie to the left of a vertical line corresponding to a specified damping factor. A necessary and sufficient condition for the set of plants in equation (1) to be simultaneously relatively stabilizable with a single control law is that the eigen values of the closed-loop system lie in the left-hand side of a vertical line in the complex s -plane. This condition motivates the following approach for determining the parameters of the PSS.

Select the parameters of the PSS to minimize the following objective function:

$$J = \max \{ \text{Re}(\lambda_{k,i}) + \beta \}; \quad k = 1, 2, 3, \dots, N; \quad i = 1, 2, \dots, n$$

where $\lambda_{k,i}$ is i^{th} closed loop eigen value of the k^{th} plant and β is relative stability factor. Subject to the constraints that finite bounds are placed on the stabilizer parameters. In this paper instead of N number of plants, single-machine-infinite-bus system is considered.

The objective function can be modified as,

$$J = \max \{ \text{Re}(\lambda_k) + \beta \}$$

The relative stability is determined by the value of β as shown in figure.2

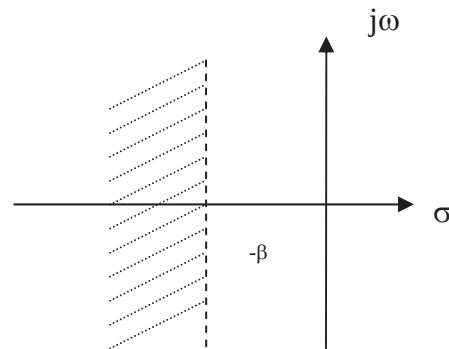


Figure 2. Region in the left-hand side of a vertical

If any solution is found such that $J < 0$, then the resulting stabilizer parameters relatively stabilize the oscillations in the collection of plants. The existence of a solution is verified numerically [5] by minimizing J .

4. System Model

In this steady single-machine-infinite-bus power system is considered [6]. The supplementary stabilizing signal considered is one proportional to speed [7]. A widely used conventional PSS is considered throughout the study. The transfer function of PSS with single phase compensation block is



$$\frac{\Delta V_s}{\Delta \omega_r} = K_{STAB} \frac{sT_w}{1+sT_w} \frac{1+sT_1}{1+sT_2}$$

The first term is stabilizer gain. The second term is washout term with a time lag T_w . The third term is a lead compensation [4] to improve the phase lag through the system.

The numerical values of T_w , T_2 and system data are given in Appendix I. The remaining parameters namely K_{STAB} and T_1 are assumed to be adjustable parameters. The optimization problem is selection of these PSS parameters easily and accurately. The optimization problem can be solved using the simulated annealing as well as particle swarm optimization. The SA algorithm is explained in Appendix II and PSO algorithm is explained in Appendix III.

For a given operating point, the power system is linearized around the operating point, the eigen values of the closed-loop system are computed, and the objective function is evaluated. It is worth mentioning that only the system electromechanical modes are incorporated in the objective function. The bounds on the parameters used in the SA are given in Appendix I.

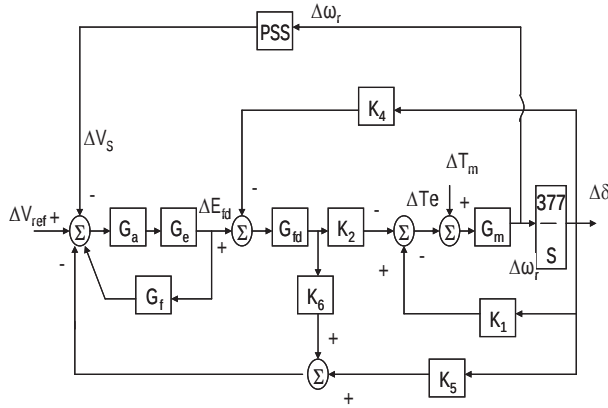


Figure 3. Block diagram of AVR with PSS

5. Simulated Annealing

Simulated Annealing [8] is an optimization technique that simulates the physical annealing process in the field of combinatorial optimization. Annealing is the physical process which involves heating up of a solid until it melts, followed by slow cooling down by decreasing temperature. Thermal equilibrium at any temperature T i.e temperature of solid is maintained as constant for a period of time, is given by Boltzmann distribution. It gives the probability of the solid being in an energy state i with energy E_i at temperature T as,

$$P_i = k e^{\left(\frac{-E_i}{T}\right)}$$

where k is Boltzmann's constant. The analogy between a physical annealing process and a combinatorial optimization problem is based on the following,

- Solutions obtained for optimization problem are equivalent to configurations of a physical system
- The cost of a solution is equivalent to the energy of a configuration.

A parameter C_p is introduced to control the temperature T . The basic elements of SA are

Trial, Current and Best Solutions:

x_{trial} , $x_{current}$, and x_{best} are set of the optimized parameter values at any iteration

Acceptance Criterion:

At any iteration, the trial solution can be accepted as the current solution if it meets one of the following criteria

- (a) $J(x_{trial}) < J(x_{current})$
- (b) $J(x_{trial}) > J(x_{current})$
- (c) $\exp\left(-\left(J(x_{trial}) - J(x_{current})\right) / C_p\right) \geq \text{rand}(0,1)$

Here $J(x_{trial})$ and $J(x_{current})$ are the values of objective function corresponding to x_{trial} and $x_{current}$ respectively. And $\text{rand}(0,1)$ is a set of random number with domain $(0,1)$.

Acceptance Ratio:

At a given value of C_p , an n_1 trial solutions can be randomly generated. Based on the acceptance criterion, an n_2 of these solutions can be accepted. The acceptance ratio is defined as n_2/n_1 .

Cooling Schedule:

It specifies a set of parameters that governs the convergence of the algorithm. This set includes an initial value of control parameter C_{p0} , a decrement function for decreasing the value of C_p , and a finite number of interactions or transitions at each value of C_p i.e. the length of each homogeneous Markov Chain. The initial value of C_p should be large enough to allow virtually all transitions to be accepted. However, this can be achieved by starting off at a small value of C_{p0} and multiplying it with a constant α larger than 1, i.e. $C_{p0} = \alpha C_{p0}$. This process continues until the acceptance ratio is close to 1. This is equivalent to heating up process in physical systems. The decrement function for decreasing the value of C_p is given by $C_p = \mu C_p$ where μ is a constant close to 1. Typical values lie between 0.8 – 0.9 [8].

Stopping Criteria:

These are the conditions under which the search process will terminate. In this study the search will terminate if any one of the following criteria is satisfied,

- (a) The number of Markov Chains since the last change of the best solution is greater than a prespecified number
- (b) The number of Markov Chains reaches the maximum allowable number.

6. Particle Swarm Optimization

Similar to evolutionary algorithms, the PSO technique conducts searches using a population of particles, corresponding to individuals. In a PSO system [9], particles change their positions by flying around in a multidimensional search space until a relatively changed position has been encountered, or until computational limitations are exceeded.

The following are the advantages of PSO [10] over other traditional optimization techniques like classical cost function and genetic algorithms,

- a) PSO is a population based search algorithm. This property ensures it to be less susceptible to getting trapped on local minima.



- b) PSO uses payoff (performance index or objective function) information to guide the search in the problem space.
- c) PSO uses probabilistic transition rules and not deterministic rules. Hence, PSO is a kind of stochastic optimization algorithm that can search a complicated and uncertain area.

Particle $X(t)$:

It is a candidate solution presented by an m-dimensional real-valued vector, where m is the number of optimized parameters.

Population $pop(t)$:

It is a set of n particles at time t. (i.e. $pop(t) = [X_1(t), X_2(t), \dots, X_n(t)]^T$)

Swarm:

It is disorganized population of moving particles that tend to cluster together while each particle seems to be moving in a random direction.

Particle velocity $V(t)$:

It is the velocity of the moving particles represented by an m-dimensional real valued vector. At the time t, the jth particle velocity $V_j(t) = [v_{j,1}(t), v_{j,2}(t), \dots, v_{j,m}(t)]$, where $v_{j,k}(t)$ is the velocity component of jth particle with respect to the kth dimension.

Individual best $X^*(t)$:

As the particle moves through the search space, it compares its fitness value at the current position to the best fitness value it has ever attained at any time up to the current time. The best position that is associated with the best fitness encountered so far is called is called individual best $X^*(t)$. For each particle in the swarm, $X^*(t)$ can be determined and updated during the search. In a minimization problem objective function J, the individual best of the jth particle $X_j^*(t)$ is determined so that $J(X_j^*(t)) \leq J(X_j^*(\tau))$, $\tau \leq t$. For simplicity, assume that $J_j^* = J(X_j^*(t))$. For the jth particle, individual best can be expressed as $X_j(t) = [x_{j,1}^*(t), x_{j,2}^*(t), \dots, x_{j,m}^*(t)]$.

Global best $X^{**}(t)$:

It is the best position among all of the individual best positions achieved so far. Hence, the global best can be determined such that $J(X^{**}(t)) \leq J(X_j^*(t))$, $j = 1, 2, \dots, n$. For simplicity, assume that $J^{**} = J(X^{**}(t))$.

Stopping criteria:

These are the conditions under which the search process will terminate. In this study, the search will terminate if one of the following criteria is satisfied,

- a) The number of iterations since the last change of the best solution is greater than a prespecified number
- b) The number of iterations reaches the maximum allowable number.

The particle velocity in the kth dimension is limited by some maximum value, $v_k^{\max}$. This limit enhances the local exploration of the problem space and it realistically simulates the incremental changes in human learning. The maximum velocity in the kth dimension is characterized by the range of the kth optimized parameter and given by

$$v_k^{\max} = \frac{(x_k^{\max} - x_k^{\min})}{N}$$

where N is a chosen number of intervals in the kth dimension.

7. Simulation Results

In this part of the study, a single machine is connected to infinite bus through a transmission line, and operating at different loading conditions [11], is considered. The linearized model of this system, voltage regulator and exciter included [6] is considered.

The constants K_1 to K_6 , with the exception of K_3 , which is only a function of the ratio of impedance, are dependent on the actual real power (P) and reactive power (Q) loading as well as the excitation levels in the machine.

The operating points are selected based on the different loading conditions. The simultaneous damping enhancement of the is demonstrated by considering five different loading conditions.

The operating points were selected randomly as follows:
 $(P_o, Q_o) = (0.9, 0.3); (0.8, -0.1); (0.5, 0.5); (0.6, -0.2); (1.0, 0.6)$

The eigen values are found by transferring the transfer function of the system data into state model.

The eigen values of the system at the five operating points considered, with out PSS are,

1. $0.4981 \pm 6.6288i, -33.6805, -17.3597$
2. $0.7513 \pm 7.3702i, -11.5526, -39.9942$
3. $0.0283 \pm 5.3580i, -25.0504 \pm 9.1822i$
4. $0.1936 \pm 6.9157i, -10.7786, -39.6528$
5. $0.5410 \pm 6.1171i, -21.6341, -29.4919$

From the eigen values it is clear that the system is unstable due to its location in the right half of the s – plane.

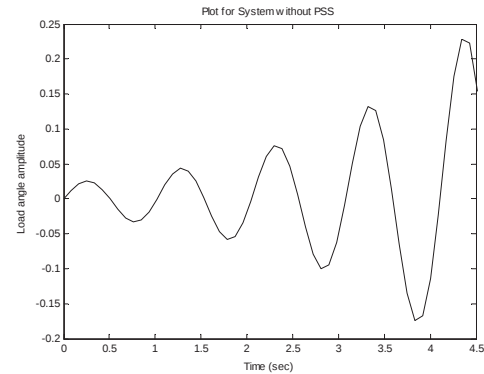


Figure 4. Response of the system without PSS

For the system to be stable the real part of eigen value should be located in the in the left hand side of s-plane.

Results for Simulated Annealing:

Simulated annealing is used to optimize the objective function J for five different kind of operating conditions to shift the eigen values to the left vertical line.

The eigen values of the five systems, with PSS are

1. $-46.3181, -16.8295 \pm 43.2325i, -0.1883 \pm 2.8091i, -0.7077$
2. $-45.7209, -16.4755 \pm 28.1318i, -0.8262 \pm 3.8261i, -0.7372$
3. $-35.6015, -21.6986 \pm 13.9605i, -0.6626 \pm 4.9864i, -0.7376$
4. $-44.0326, -16.5329 \pm 18.3720i, -1.6023 \pm 4.3745i, -0.7585$
5. $-45.6105, -17.2183 \pm 40.4834i, -0.1521 \pm 2.8783i, -0.7102$



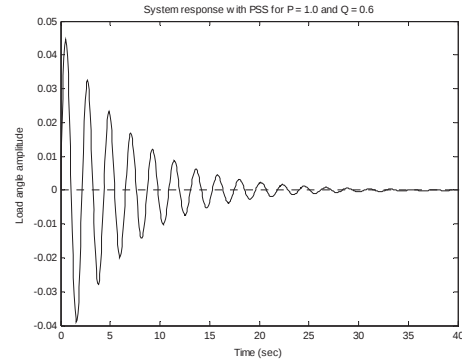
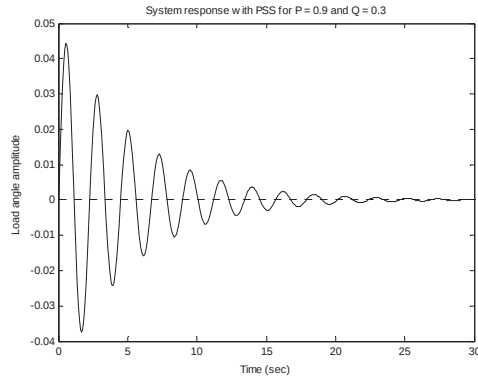


Figure 5. Response of the system with PSS using Simulated Annealing

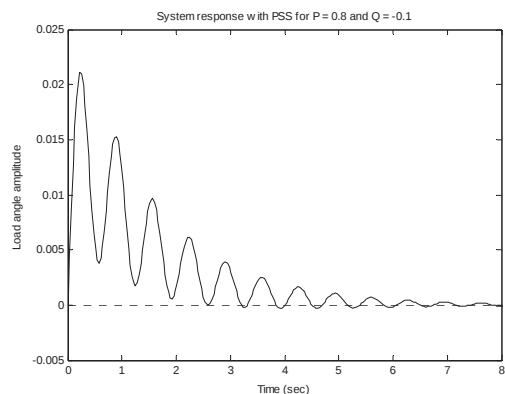
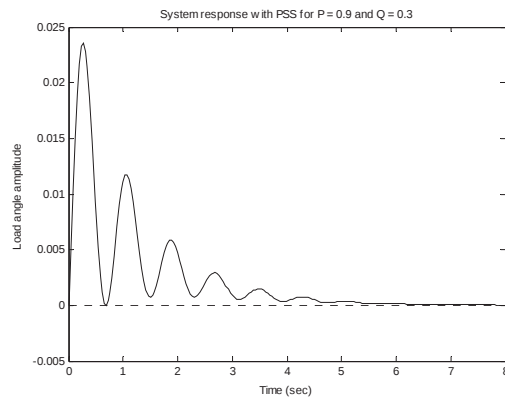
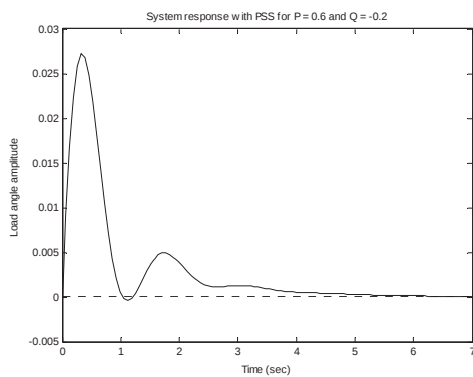
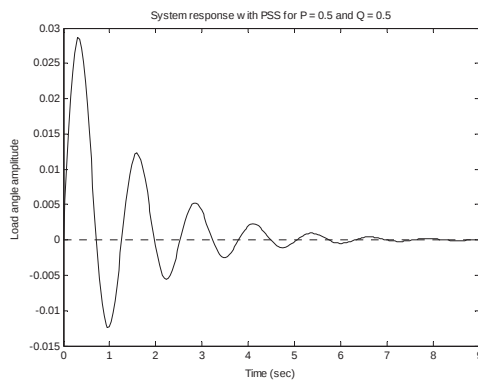
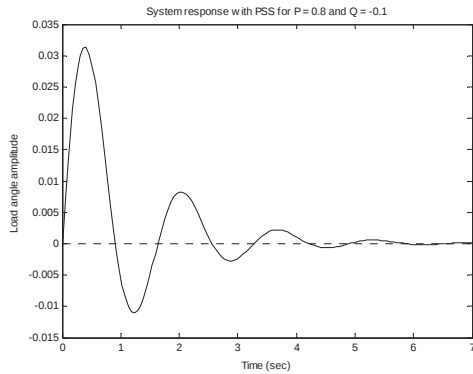
From figure.5 it is obvious that the rotor oscillations are damped and the above results shows that all the eigen values of the system were located in the left half of the s – plane to make the system stable.

Results for Particle Swarm Optimization:

The objective function J is optimized with the PSO and $N=5$ to shift the electromechanical mode of each of the five systems to the left of the vertical line defined by $\beta = -2.2$.

The eigen values of the five systems, with PSS are

1. $-34.1670 \pm 5.0796i, -9.9855, -0.9887 \pm 7.7731i, -0.7644$
 2. $-37.4972, -35.6751, -0.5533 \pm 9.3491i, -6.0033, -0.7792$
 3. $-29.0306 \pm 7.0224i, -20.4992, -0.8755 \pm 5.6109i, -0.7500$
 4. $-36.4206 \pm 0.9044i, -1.3020 \pm 9.2146i, -4.8165, -0.7998$
 5. $-33.0536 \pm 5.8900i, -12.5555, -0.8183 \pm 6.9006i, -0.7620$
- All the eigen values of the system were located in the left half of the s – plane. So the system will be stable.



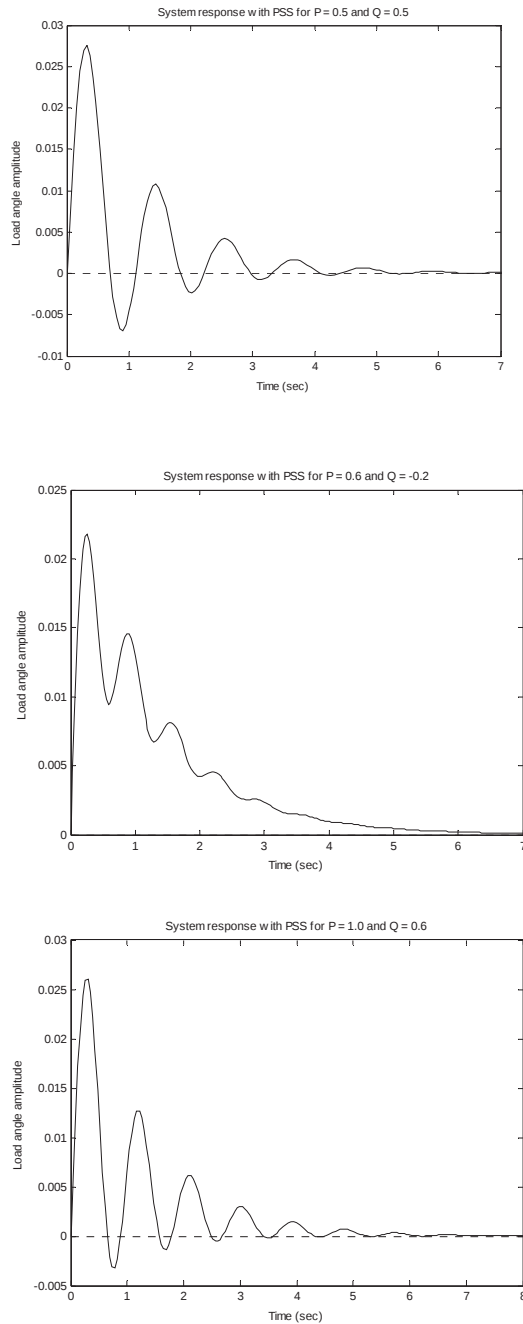


Figure 6. Response of the system with PSS using Particle Swarm Optimization

Figure.6 shows response of the system with power system stabilizer for various operating conditions whereas their parameters were tuned by using Particle Swarm Optimization (PSO). It is indicating simultaneous improvement in the response of the five systems. The above two techniques can be compared as shown in Table 1,

Table 1: Comparison between simulated annealing and Particle Swarm Optimization

Loading Condition (P_o, Q_o)	Simulated Annealing		Particle Swarm Optimization	
	Settling Time in Sec.	Peak Amp.	Settling Time in Sec.	Peak Amp.
0.9, 0.3	20.8	0.0443	4.51	0.0236
0.8, -0.1	4.6	0.0313	6.32	0.0211
0.5, 0.5	5.58	0.0287	4.92	0.0275
0.6, -0.2	3.96	0.0272	5.06	0.0218
1.0, 0.6	4.09	0.0324	4.99	0.026

From the above comparison the simulated annealing is having higher settling time, higher peak amplitude and higher computational time than Particle swarm optimization. So tuning of PSS parameters by using PSO is more optimal than SA, Classical Cost Function and Genetic Algorithm [12]. If we considered N plants, the same results hold good for individual plants.

8. Conclusion

The use of Simulated Annealing and PSO to design robust power system stabilizers for power systems working at various operating conditions are investigated in this paper. The problem of selecting the PSS parameters, which simultaneously improve the damping at various operating conditions, is converted to an optimization problem with an eigen value-based objective function which is solved by both Simulated Annealing and PSO. For a general loading condition with real power 0.9 and reactive power 0.3, the simulated annealing will settle down oscillations of the system under study at 20.8 seconds with peak amplitude of 0.0443, whereas PSO will settle down oscillations at 4.51 seconds with peak amplitude of 0.0236. So it is evident that PSO is more optimal than SA for this case.

The objective function is presented allowing the robust selection of stabilizer parameters that will optimally place the closed-loop eigen values in the left hand side of a vertical line in complex s-plane. By comparing the above two meta-heuristic optimization techniques, it is found that Particle Swarm Optimization is better than simulated annealing in tuning the parameters of the Power system stabilizer, to reduce intra and inter area rotor oscillations over a wide range of operating conditions.



Appendix – I

System Data

Single- Machine- Infinite- Bus System:

$$G_m = \frac{1}{2H_s + K_D} ; \quad G_A = \frac{K_A}{sT_A + 1} ; \quad G_e = \frac{1}{sT_E + 1} ;$$

$$G_f = \frac{sK_f}{1 + sT_F} ; \quad G_{fd} = \frac{K_3}{1 + sT_{do}K_3}$$

The system data are as follows:

Machine (p.u)

$$x_d = 1.7 ; x'_d = 0.254 ; x_q = 1.64 ; \omega_0 = 120 \pi \text{ rad/s} ;$$

$$T'_{do} = 5.9 \text{ sec} ; K_D = 0 ; H = 2.37 \text{ sec}$$

Transmission Line (p.u)

$$r_e = 0.02 ; x_e = 0.4$$

Exciter and Stabilizer

$$K_A = 400 ; T_A = 0.05 \text{ sec} ; K_F = 0.025$$

$$T_F = 1.0 \text{ sec} ; K_E = -0.17 ; T_E = 0.95 \text{ sec}$$

$$T_W = 10 \text{ sec} ; T_2 = 0.0227 \text{ sec}$$

Bounds for the stabilizer adjustable gain and time constants are [0.01, 10] and [0.03, 1.0] respectively.

APPENDIX – II

SIMULATED ANNEALING ALGORITHM

- Step (1): Set the initial value of C_{p0} and randomly generate an initial solution x_{initial} and calculate its objective function. Set this solution as the current solution as well as best solution, i.e. $x_{\text{initial}} = x_{\text{current}} = x_{\text{best}}$.
- Step (2): Randomly generate an n_1 of trial solutions in the neighborhood of the current solution.
- Step (3): Check acceptance criterion of these trial solutions and calculate the acceptance ratio. If the acceptance ratio is close to 1 go to step 4, else set $C_{p0} = \alpha C_{p0}$, $\alpha > 1$, and go back to step 2.
- Step (4): Set the chain counter $k_{\text{ch}} = 0$.
- Step (5): Generate a trial solution x_{trial} . If x_{trial} satisfies the acceptance criterion, set $x_{\text{current}} = x_{\text{trial}}$, $J(x_{\text{current}}) = J(x_{\text{trial}})$, and go to step 6; else go to step 6.
- Step (6): Check the equilibrium condition. If it is satisfied go to step 7, else go to step 5.
- Step (7): Check the stopping criteria. If one of them is satisfied then stop; else set $k_{\text{ch}} = k_{\text{ch}} + 1$ and $C_p = \mu C_p$, $\mu < 1$, and go back to step 5.

APPENDIX – III

PARTICLE SWARM OPTIMIZATION

Step (1): Set the time counter $t = 0$ and generate random n particles, $\{X_j(0), j = 1, 2, \dots, n\}$, where $X_j(0) = [x_{j,1}(0), x_{j,2}(0), \dots, x_{j,m}(0)]$. $x_{j,k}(0)$ is generated by randomly selecting a value with uniform probability over the k th optimized parameter search space $[x_k^{\min}, x_k^{\max}]$.

Generate randomly initial velocities of all particles, $\{V_j(0), j = 1, 2, \dots, n\}$, where $V_j(0) = [v_{j,1}(0), v_{j,2}(0), \dots, v_{j,m}(0)]$. $v_{j,k}(0)$ is generated by randomly selecting a value with uniform probability over the k th optimized parameter search space $[-v_k^{\max}, v_k^{\max}]$.

Each particle in the initial population is evaluated using the objective function, J . For each particle, set $X_j^*(0) = X_j(0)$ and $J_j^* = J_j$, $j=1, 2, \dots, n$. Search for the best value of objective function J_{best} . Set the particle associated with J_{best} as the global best, $X^{**}(0)$, with an objective function of J^{**} . Set the initial value of the inertia weight $w(0)$.

Step (2): Update the time counter $t = t + 1$.

Step (3): Update the inertia weight $w(t) = \alpha w(t-1)$.

Step (4): Using the global best and individual best, the j th particle velocity in the k th dimension is updated according to the following equation:

$$v_{j,k}(t) = w(t)v_{j,k}(t-1) + c_1r_1(x_{j,k}^*(t-1) - x_{j,k}(t-1)) + c_2r_2(x_{j,k}^{**}(t-1) - x_{j,k}(t-1))$$

where c_1 and c_2 are positive constants and r_1 and r_2 are uniformly distributed random numbers in $[0, 1]$.

Step (5): Based on the updated velocities, each particle changes its position according to the following equation:

$$x_{j,k}(t) = v_{j,k}(t) + x_{j,k}(t-1)$$

Step (6): Each particle is evaluated according to the updated position. $J_j < J_j^*$, $j=1, 2, \dots, n$, then update individual best as $X_j^*(t) = X_j(t)$ and $J_j^* = J_j$, and go to step 7; else go to step 7.

Step (7): Search for the minimum value $J_{\min}$ among J_j^* , where $\min$ is the index of the particle with minimum objective function value, i.e., $\min \in \{j; j=1, 2, \dots, n\}$. If $J_{\min} < J^{**}$ then update global best as $X^{**} = X_{\min}(t)$, and $J^{**} = J_{\min}$ and go to step 8; else go to step 8.

Step (8): If one of the stopping criteria is satisfied, then stop, or else go to step 2.

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Direct Adaptive Fuzzy Control of Nonlinear Systems

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Abstract

This paper proposes robust adaptive fuzzy controller for unknown nonlinear systems with disturbances. The control law is composed of primary and auxiliary control terms. The first term is based on Takagi-Sugeno fuzzy approach with feedback linearization method. The second term is obtained by combining the Sliding Mode Control (SMC) and Lyapunov approach to compensate the unmodeled dynamics and disturbance effects. The stability of the system has been achieved considering a Lyapunov approach. The capability of the proposed method is illustrated through two numerical examples.

Keywords: Nonlinear systems, Adaptive control, Fuzzy logic, Sliding mode.

1. Introduction

In the last years, several control methods of nonlinear systems have appeared in the literature [1, 2, 4]. Feedback linearization is suitable for the control of nonlinear systems with accurate nominal models or nonlinearly parametrical dynamical model. On the other hand, the LMI approach is combined with H_∞ technique for ensuring the stability and the robustness. The SMC method is proposed to provide robust controller systems. However, SMC suffers from a well known problem, chattering due to the high gain and high speed switching control. One of the common solutions for eliminating this chattering effect is to introduce a boundary layer neighbouring of the sliding surface [4]. However, these approaches are difficult to implement and their use requires restrictive conditions; the dynamics of system are assumed to be known or linear with a set of unknown parameters. In fact, most of the parameters and structure are unknown due to environment changes, modelling error and unmodeled dynamics. Face the above problems; there exist several techniques, in particular; the intelligent technologies as neural networks, fuzzy logic, genetics algorithms and evolutive computation [3, 9, 10]. Fuzzy controllers are supposed to work in situations where there is a large uncertainty or unknown variation in plant parameters and structure [6, 8, 12]. Practical implementation of these algorithms is difficult when the

system presents strongly nonlinearity and perturbed. Fuzzy control has been successfully applied to many industrial systems where no accurate mathematical models of the systems under control are available and human experts are available to provide linguistic fuzzy control rules or linguistic fuzzy descriptions about the systems [7, 13, 14]. The apparent similarities between sliding mode control and fuzzy control motivate considerable research efforts in combining the two approaches for achieving more superior performances such as overcoming some limitations of the traditional sliding mode control [3, 5, 11].

In this paper, a robust adaptive control is proposed for a class of unknown nonlinear system in presence of the disturbances. The strategy of the control is based on combining the fuzzy approach and SMC. The T-S adaptive fuzzy technique is used to approximate the nonlinear dynamics and disturbances. By this approximation, we can transform the nonlinear problem to a linear one and then the identification and synthesis of the primary control law become more performing and easy. The SMC is introduced in order to ensure the robustness in presence of the disturbances and unmodeled dynamics. The controller proposed in this paper is composed of two terms; an auxiliary control law which is designed to compensate the effect of unmodeled dynamics, disturbances and uncertainties of parameters and a primary control law which ensure a good tracking. Another advantage of this controller is the using of Lyapunov approach to guarantee the stability and convergence analysis. The parameters of this controller are directly estimated which reduce the computational load and ensure a rapid parameter convergence.

The paper is organised as follows: The problem formulation is presented in Section 2. Section 3, gives the direct adaptive fuzzy control design. Simulation results are presented in Section 4.

2. Problem Formulation

Consider a class of nonlinear systems described as follows:



$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = x_3 \\ \vdots \\ \dot{x}_n = f(\underline{x}, t) + b u(t) + d(t) \\ y = x_1 \end{cases} \quad (1)$$

Or equivalently:

$$\begin{cases} \dot{x}^{(n)} = f(\underline{x}, t) + b u(t) + d(t) \\ y = x \end{cases} \quad (2)$$

where $\underline{x} = [x, \dot{x}, \dots, x^{(n-1)}]^T = [x_1, x_2, \dots, x_n]^T \in R^n$ is the state vector of the system.

$u \in R$ and $y \in R$ are respectively the input and output of the system. $d(t)$ represents the unknown external disturbances.

We have to make the following assumptions:

A.1 The controllability is ensured if the constant b is strictly positive.

A.2 The observability is ensured since the state vector $\underline{x}$ is assumed to be available for measurement.

The objective is to design an adaptive control which will cause the output y to track a desired reference y_r (i.e., $e_0 \rightarrow 0$) and have up to $(n-1)$ bounded derivatives.

The tracking error is defined as:

$$e_0 = y - y_r \quad (3)$$

The reference vector is denoted by:

$$Y_r = [y_r, \dot{y}_r, \dots, y_r^{(n-1)}]^T$$

Then the error vector is given by:

$$\underline{e} = [e_0, \dot{e}_0, \dots, e_0^{(n-1)}]^T \in R^n$$

We note that the stability of the closed-loop system is guaranteed in the sense that the input and the output signals are bounded.

We remark that, when the system (2) is well known without disturbances, the primary control should be designed to have the following control law [2]:

$$u^* = \frac{1}{b} (-f(\underline{x}, t) + y_r^{(n)} - \sum_{i=1}^{n-1} k_i e_0^{(i)}) \quad (4)$$

The selection of $k_i; i=1, 2, \dots, n-1$ must satisfy the following Hurwitz polynomial:

$$H(z) = z^{n-1} + k_{n-1}z^{n-2} + \dots + k_1$$

In fact, the nonlinear function $f(\underline{x}, t)$ is unknown and composed by nonlinear dynamic and disturbances. We note that any restrictive conditions are made about the nonlinearity. In these circumstances, the primary control law (4) becomes very difficult to implement and not always an evident task. The proposed strategy consists to transform the nonlinear problem to a linear one by using the T-S fuzzy model and combining the Lyapunov approach with SMC technique to derive a robust control.

3. Direct adaptive fuzzy control

In this section, the direct adaptive fuzzy control is designed for a class of nonlinear system with unknown nonlinear dynamics and disturbances. The strategy of

control is based on fuzzy logic and SMC approach to ensure stability and good tracking.

At first level, the primary control law (4), is approximated by the adaptive Takagi-Sugeno (T-S) fuzzy model. The parameters of the controller are tuned on-line from Lyapunov approach. In the second level, the auxiliary part of control is added to compensate the external disturbances and the unmodeled dynamics.

The fuzzy logic controller includes three important steps: fuzzification, fuzzy inference and defuzzification. The fuzzification comprises the process of transforming crisp values into grades of membership for linguistic terms of fuzzy sets. The membership function is used to associate a grade to each linguistic term. The basic operation of the inference process is to determine the values of the controller output based on the contributions of each rule in rule base. The defuzzification is the process of producing a quantifiable result in fuzzy logic.

The fuzzy controller is described by a collection fuzzy rule if-then as follows:

$$R^l: \text{If } x_1 \text{ is } F_1^l \text{ and } x_2 \text{ is } F_2^l \dots \text{and } x_n \text{ is } F_n^l \text{ then } u^l \text{ is } \theta_l$$

$$l = 1, 2, \dots, m$$

where R^l ($1 \leq l \leq m$) denotes the l th implication, F_i^l ($1 \leq i \leq n$) are fuzzy variables characterized by membership functions $\mu_{F_i^l}$ ($1 \leq i \leq n$).

By using the product inference and the center of gravity COG method [13], the crisp value of the output is obtained using the following relation:

$$\hat{u}(\underline{x}, \underline{\theta}) = \frac{\sum_{l=1}^m \theta_l \prod_{i=1}^n \mu_{F_i^l}}{\sum_{l=1}^m \prod_{i=1}^n \mu_{F_i^l}} = \underline{\theta}^T \underline{\xi}(\underline{x}) \quad (5)$$

where m is the total number of fuzzy rules and $\underline{\theta} = [\theta_1, \theta_2, \dots, \theta_m]^T$ is the adjustable parameter vector grouping all consequent of the controller.

$\underline{\xi}(\underline{x}) = [\xi_1(\underline{x}), \xi_2(\underline{x}), \dots, \xi_m(\underline{x})]^T$ is the vector of the fuzzy basis functions

$$\xi_l(\underline{x}) = \frac{\prod_{i=1}^n \mu_{F_i^l}}{\sum_{l=1}^m \prod_{i=1}^n \mu_{F_i^l}} \quad (6)$$

It is nontrivial to apply directly the control law (5) due to the existence of the unmodeled dynamics and the disturbances. For achieving a good tracking, an auxiliary signal is incorporated in the control law (5). This signal is designed to compensate the effect of the unmodeled dynamics and the disturbances. The synthesis of a robust controller is based on combining a SMC to a Lyapunov approach. The SMC requires that bounds on the modelling uncertainties and disturbances are known in order to provide robust stability. However, chattering in the control signal occurs because the conservative nature of the controller is highly undesirable in many systems. Chattering can be eliminated by using the boundary layer method [2].



Define a time varying sliding surface $\sigma(\underline{x}, t) = 0$, in the state space R^n by the scalar equation:

$$\sigma(\underline{x}, t) = e_0^{(n-1)} + k_{n-1}e_0^{(n-2)} + \dots + k_1e_0 \quad (7)$$

Define the optimal parameter vectors as:

$$\underline{\theta}^* = \arg \min_{\underline{\theta} \in \Omega} \left(\sup_{\underline{x} \in R^n} |u^* - \hat{u}(\underline{x}, \underline{\theta})| \right) \quad (8)$$

where $\Omega = \{\underline{\theta} / |\underline{\theta}| \leq M\}$ is the convex compact set with M is a pre-specified parameter.

The parameter error is defined as

$$\Phi = \underline{\theta} - \underline{\theta}^* \quad (9)$$

$$\dot{\sigma} = y^{(n)} - y_r^{(n)} + \sum_{i=1}^{n-1} k_i e_0^{(i)} \quad (10)$$

$$= b(\hat{u}(\underline{x}, \underline{\theta}) + u_s + d(t) - u^*)$$

Let the minimum approximation error as follows:

$$w = u(\underline{x}, \underline{\theta}^*) - u^* \quad (11)$$

$$\begin{aligned} \dot{\sigma} &= b(\hat{u}(\underline{x}, \underline{\theta}) - \hat{u}(\underline{x}, \underline{\theta}^*) + u_s + d(t) + \hat{u}(\underline{x}, \underline{\theta}^*) - u^*) \\ &= b(\Phi^T \xi(\underline{x}) + u_s + d(t) + \hat{u}(\underline{x}, \underline{\theta}^*) - u^*) \end{aligned} \quad (12)$$

Assumptions

A.3 There exist a positive constant $w_{\max}$ such that $|\hat{u}(\underline{x}, \underline{\theta}) - u^*| < w_{\max}$.

A.4 Upper bound of the unknown external disturbances is $d_{\max}$ and satisfying the following condition: $k_d \geq w_{\max} + d_{\max}$ where k_d is a positive constant.

The direct adaptive control law becomes:

$$u = \hat{u}(\underline{x}, \underline{\theta}) + u_s \quad (13)$$

The auxiliary control is given as:

$$u_s = -k_d \text{sat}(\sigma) \quad (14)$$

The function $\text{sat}(\sigma)$ can be written as:

$$\text{sat}(\sigma) = \begin{cases} \text{sgn}(\sigma) & |\sigma| > \varepsilon \\ \frac{\sigma}{\varepsilon} & |\sigma| \leq \varepsilon \end{cases} \quad (15)$$

with ε is a positive constant. It is chosen small to vicinity of the origin.

The adjustable fuzzy parameters of $\hat{u}(\underline{x}, \underline{\theta})$ are tuned on line using the Lyapunov approach. In order to guarantee that the parameters are bounded, we introduce the projection algorithm [12] to restrict them in the closed set Ω .

$$\dot{\underline{\theta}} = \begin{cases} -\gamma \sigma b \xi(\underline{x}) & (|\underline{\theta}| < M \text{ or } (|\underline{\theta}| = M \text{ and } \gamma \sigma b \xi^T(\underline{x}) > 0)) \\ \text{Proj}[-\gamma \sigma b \xi(\underline{x})] & \text{otherwise} \end{cases} \quad (16)$$

where γ is fixed adaptive gain. $\text{Proj}[\cdot]$ represents the projection operator which is defined by:

$$\text{Proj}[-\gamma \sigma b \xi(\underline{x})] = -\gamma \sigma b \xi(\underline{x}) + \gamma \sigma b \frac{\underline{\theta} \underline{\theta}^T \xi(\underline{x})}{|\underline{\theta}|^2} \quad (17)$$

Theorem: For the system (2), if the control law (13) is used under the assumptions A.1-4, the resulting closed loop system is stable in sense that:

(i) The input and the output signals are bounded.

(ii) The tracking performance is achieved.

Proof: Consider the following Lyapunov function:

$$V = \frac{1}{2} \sigma^2 + \frac{1}{2\gamma} \Phi^T \Phi \quad (18)$$

The time derivative of V equals:

$$\begin{aligned} \dot{V} &= \sigma \dot{\sigma} + \frac{1}{\gamma} \Phi^T \dot{\Phi} \\ &= \sigma b (\Phi^T \xi(\underline{x}) + w + d(t) + u_s) + \frac{1}{\gamma} \Phi^T \dot{\Phi} \\ &= \frac{1}{\gamma} \Phi^T (\dot{\Phi} + \gamma \sigma b \xi(\underline{x})) + \sigma b (w + d(t) + u_s) \end{aligned}$$

$$\dot{V} \leq \frac{1}{\gamma} \Phi^T (\dot{\Phi} + \gamma \sigma b \xi(\underline{x})) + b|\sigma|(w_{\max} + d_{\max}) + b\sigma u_s \quad (19)$$

where $\dot{\Phi} = \dot{\underline{\theta}}$.

Substitute (17) into (19) and taken k_d with the previous assumption.

Then we have:

$$\dot{V} \leq b|\sigma|(w_{\max} + d_{\max}) + b\sigma u_s \quad (20)$$

if $|\sigma| > \varepsilon$, $\text{sat}(\sigma) = \text{sgn}(\sigma)$, then $u_s = -k_d \text{sgn}(\sigma)$

$$\dot{V} \leq b|\sigma|(w_{\max} + d_{\max}) - k_d b|\sigma|, \text{ so } \dot{V} \leq 0.$$

if $|\sigma| \leq \varepsilon$, $\text{sat}(\sigma) = \frac{\sigma}{\varepsilon}$, then $u_s = -k_d \frac{\sigma}{\varepsilon}$

$$\dot{V} \leq b|\sigma|(w_{\max} + d_{\max}) - k_d b \frac{\sigma^2}{\varepsilon}, \text{ hence } \dot{V} \leq 0.$$

4. Simulation results

In this section, we test the proposed controller algorithm on two examples. The first example is a regulation problem of a nonlinear servomechanism [6]. The second example is to let a doffing forced-oscillation system [12] to track a desired trajectory.

Example 1: The nonlinear servomechanism is modelled by the following second order differential equation:

$$m\ddot{q} + \ell\dot{q} + \Delta f(q) = \tau + d \quad (21)$$

where

- $\dot{q}$: Velocity,
- q : Position,
- $\Delta f(q)$: Nonlinear term depending on q ,
- m, ℓ : Mass and damping,
- τ : Torque,
- d : Disturbances.

We suppose that the position $x_1 = q$ and the velocity $x_2 = \dot{q}$ are available from measurements.

The dynamic equations of the servomechanism can be described in space state as:

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = f(\underline{x}, t) + u(t) + d(t) \\ y = x_1 \end{cases} \quad (22)$$

Or equivalently

$$\begin{cases} \dot{x}^{(2)} = f(\underline{x}, t) + b u(t) + d(t) \\ y = x \end{cases} \quad (23)$$

with $f(\underline{x}, t) = -\ell x_2 - \Delta f(x_1)$ and $u = \tau$.



where $\Delta f(x_1) = 0.4 \sin(x_1)$.

The control objective is to maintain the system to track the desired angle trajectory: $y_r = \frac{\pi}{3} \sin(0.01t)$.

The parameters of the system are given as:

$$m = 1 \text{ kg}, \quad l = 1 \quad \text{and} \quad d(t) = 2 \cos(0.01t).$$

The sliding surface is defined as:

$$\sigma(\underline{x}, t) = k_1 e_0(t) + k_2 \dot{e}_0(t) \quad \text{with} \quad k_1 = 5 \quad \text{and} \quad k_2 = 1.$$

We choose the initial condition as:

$$x_1(0) = 1 \quad \text{and} \quad x_2(0) = 0. \quad \text{The learning rate } \gamma = 0.5.$$

In the first step, we need to define some fuzzy sets to cover the state space. The choice of the number of fuzzy set and the constant M is related to knowledge of expert on the system. For simplicity, we consider $M = 6$ and the fuzzy membership functions (see figure 1) are chosen as:

$$\mu_{F_1^1}(x_1) = \exp(-0.5(x_1 + \pi/3));$$

$$\mu_{F_1^2}(x_1) = \exp(-0.5(x_1));$$

$$\mu_{F_1^3}(x_1) = \exp(-0.5(x_1 - \pi/3));$$

$$\mu_{F_2^1}(x_2) = \exp(-0.5(x_2 + \pi/3));$$

$$\mu_{F_2^2}(x_2) = \exp(-0.5(x_2));$$

$$\mu_{F_2^3}(x_2) = \exp(-0.5(x_2 - \pi/3));$$

(24)

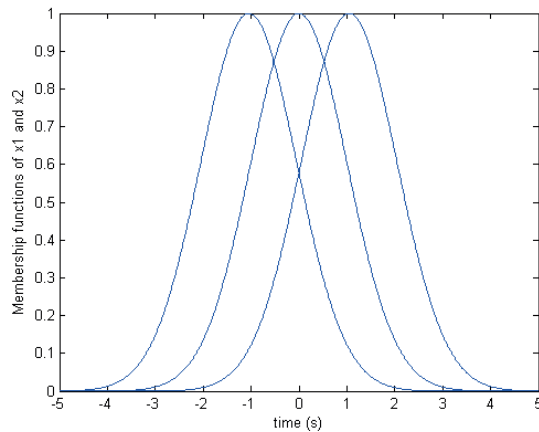


Figure 1. Membership function of x_1 and x_2

Then there are 9 rules to approximate the primary control law $\hat{u}$. The fuzzy rules are defined by the following linguistics description:

$$R^l : \text{If } x_1 \text{ is } F_1^l \text{ and } x_2 \text{ is } F_2^l \text{ then } u^l \text{ is } \theta_l$$

By using the singleton fuzzification, product inference and COG method defuzzification, the primary control is given by :

$$\hat{u}(\underline{x}, \theta) = \frac{\sum_{l=1}^9 \theta_l \prod_{i=1}^2 \mu_{F_i^l}}{\sum_{l=1}^9 \prod_{i=1}^2 \mu_{F_i^l}} = \underline{\theta}^T \underline{\xi}(\underline{x})$$

The auxiliary gain control is chosen as: $k_d = 3$.

We have 9 parameters of controller to estimate, their convergence is given on figure 2.

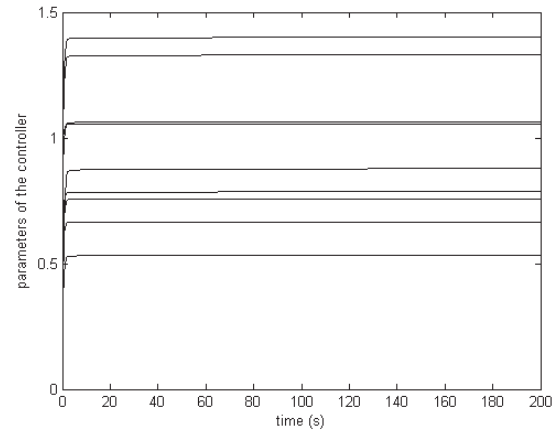


Figure 2. Evolution of parameters of the primary controller

The behaviour of $y(t)$ and $y_r(t)$ is given on figure 3, it can be seen that a good tracking is obtained in presence of nonlinearity and disturbances.

The corresponding fuzzy control signal is given on figure 4.

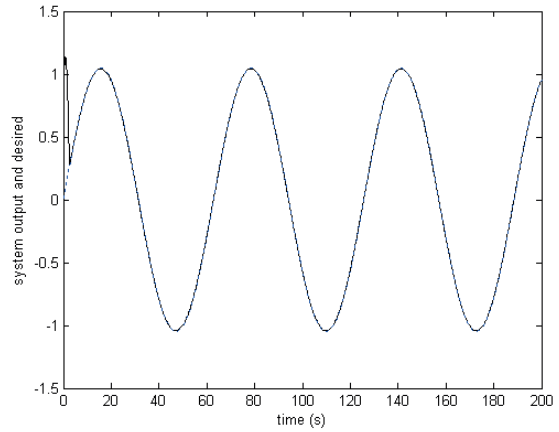


Figure 3. Responses of $y(t)$ and $y_r(t)$

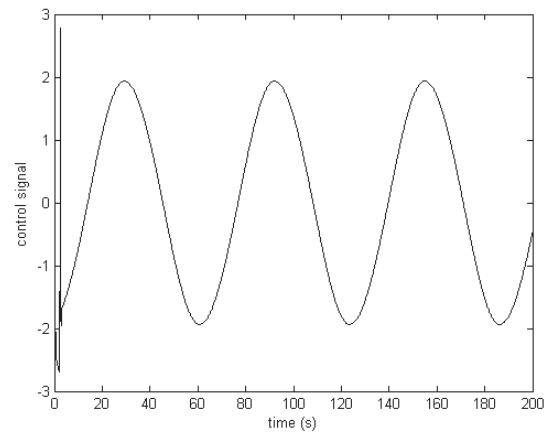


Figure 4. The control signal

Figures 5 and 6 gives respectively the convergence of the error and the sliding surface.

Example 2: Consider a doffing forced-oscillation system described by:



$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = -0.1x_2 - x_1^3 + 12\cos(t) + bu + d(t) \\ y = x_1 \end{cases} \quad (25)$$

Or equivalently:

$$\begin{cases} \dot{x}^{(2)} = -0.1x_2 - x_1^3 + 12\cos(t) + bu(t) + d(t) \\ y = x \end{cases} \quad (26)$$

The desired angle trajectory is chosen as follows: $y_r(t) = \sin(0.015t)$. We choose the sliding surface as:

$$\sigma = k_1 e_0(t) + k_2 \dot{e}_0(t) \text{ with } k_1 = 0.5 \text{ and } k_2 = 0.1.$$

Let the initial values: $x_1(0) = 1, x_2(0) = 1$, the learning rate $\gamma = 0.05, M = 6, k_d = 15$ and step size $0.01s$.

The external disturbances are given by $d(t) = 0.1rand(1,20)$. To construct the fuzzy controller $\hat{u}(x, \theta)$, we define fuzzy sets for component of each x_1 and x_2 . The membership functions are represented by figures 7 and 8. There are 14 rules to approximate $\hat{u}(x, \theta)$. There are 14 parameters of controller to estimate their evolution is given on figure 9. We remark that a good convergence is obtained. The tracking of $y(t)$ is presented on figure 10, it is similar to that on figure 3. The corresponding fuzzy control signal is shown in figure 11.

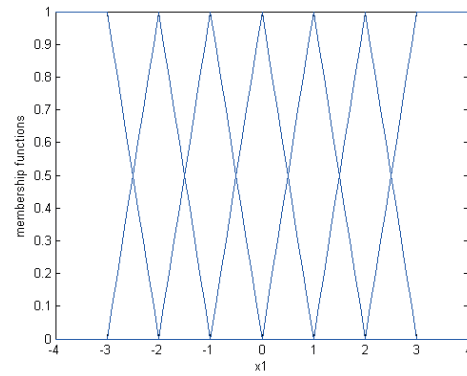


Figure 7. Membership functions of x_1

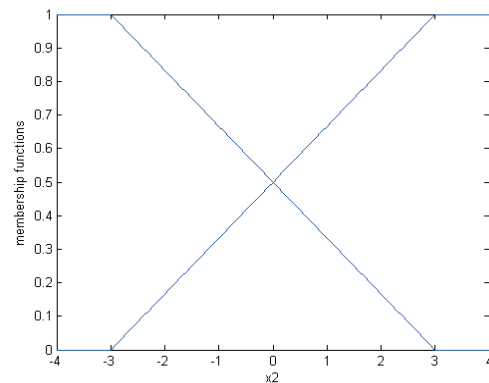


Figure 8. Membership functions of x_2

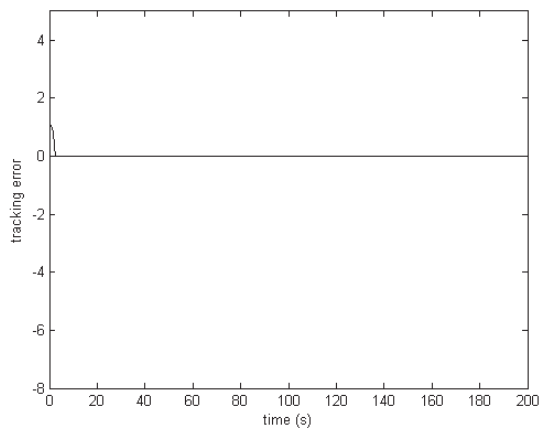


Figure 5. The behaviour of the error $e(t)$

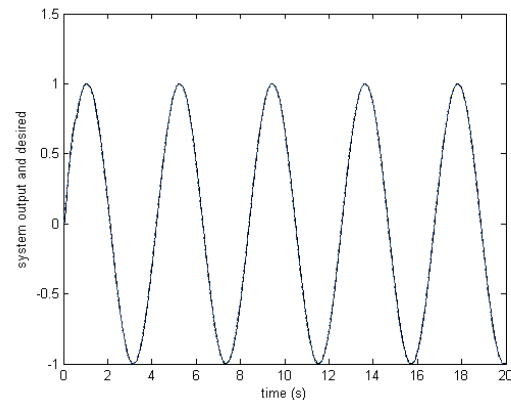


Figure 9. Responses of $y(t)$ and $y_r(t)$

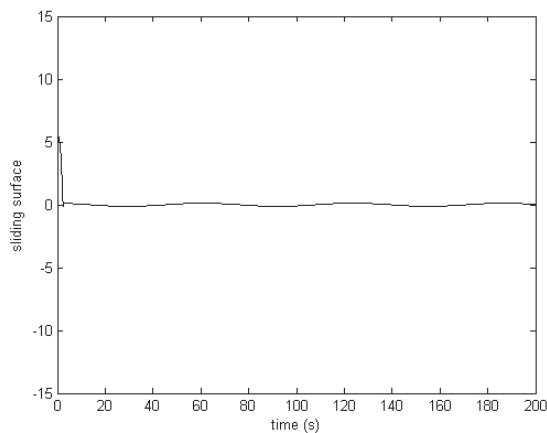


Figure 6. The sliding surface

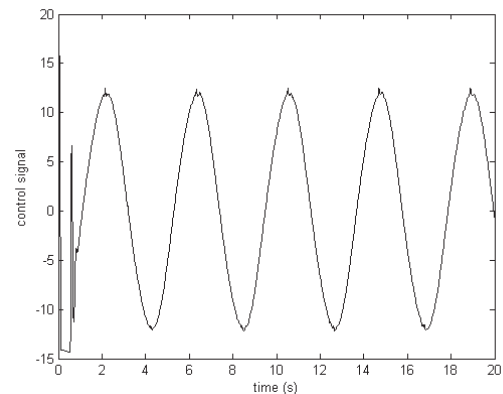


Figure 10. The control signal



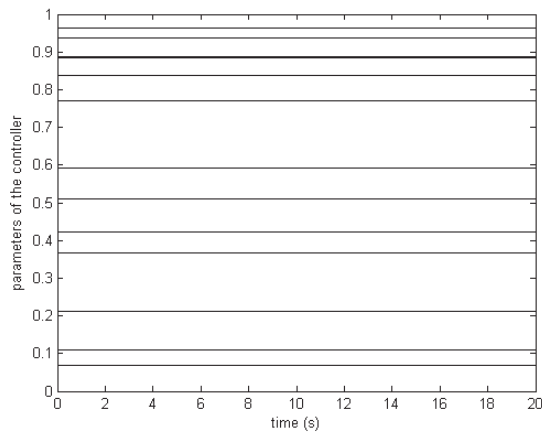


Figure11. Evolution of parameters of the primary controller

Figures 12 and 13, shows respectively the convergence of the error and the sliding surface.

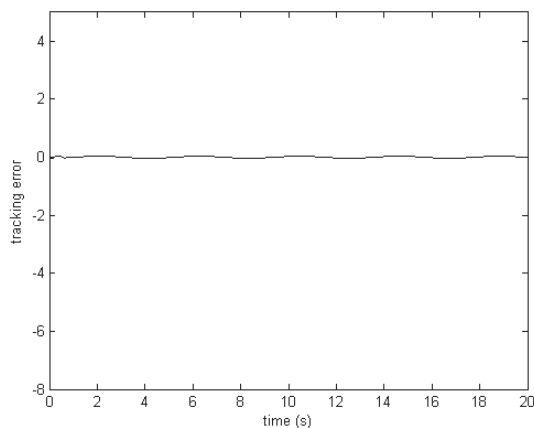
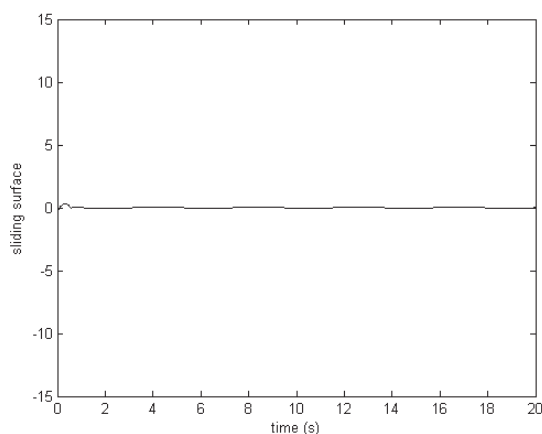
Figure12. The behaviour of the error $e(t)$ 

Figure 13. The sliding surface

5. Conclusion

In this paper, we have presented a direct adaptive fuzzy controller for a class of nonlinear systems with unknown dynamics and disturbances. The synthesis of the direct adaptive method is based on fuzzy logic to estimate the primary control. The unmodeled dynamics and disturbances are taken into account in the control law by incorporating an auxiliary signal. This signal is determined by combining the SMC and Lyapunov approach which allows to synthesise a robust controller. It is shown on numerical examples that the proposed approach is really a robust method and can deal with a large class of nonlinear systems.

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A Wireless Sensor-Based Driving Assistant for Automobiles

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Abstract

This paper presents a new concept and an experimental validation of a real-time driver assistance system involving multiple sensors and actuators monitoring an area around the vehicle and conveying alerts to the driver. The proposed system, dubbed the driving assistant, detects the presence of obstacles within the monitored area and alerts the driver via a combination of tactile, audio, and visual signals. It features simple ultrasonic sensors installed at the two front corners and the two blind spots of the vehicle. Our design is inexpensive and flexible, which makes it deployable as an add-on to existing vehicles. In particular, the interconnection of modules is achieved via a low-bandwidth low-power wireless link. We explain the rationale behind the driving assistant concept, discuss its performance within the context of human reaction time affecting the safety of vehicle operation, and suggest ways of incorporating such a system into real-life cars.

Keywords: *Actuators, Real-time Control, Ultrasonic Sensors, Wireless Communication.*

1. Introduction

Every year, in the United States alone, approximately six million automobile accidents take place. In total, those accidents result in about 43,000 deaths and 2.9 million injuries, not to mention the huge financial loss amounting to over 230 billion US dollars [1].

A large fraction of all automobile accidents is caused by the drivers' lack or lapse of concentration while operating their vehicles. Some drivers tend to occupy themselves with distracting activities, such as tuning the radio, eating, talking to passengers, or making cellular phone calls. Other drivers find it difficult to maintain focus on driving, e.g., due to fatigue or health problems [2]. Elderly drivers may exhibit difficulties in personal mobility making it more difficult for them to reliably monitor the vehicle perimeter [3]. They may also develop other conditions having a negative (albeit not disqualifying) impact on their ability to focus on the road [4][5][6].

In this paper, we discuss the design concept and operation of a driving assistant, whose responsibility is to

alert the driver to the presence of potentially hazardous objects within the vehicle's perimeter. By combining tactile, audio, and visual feedback, the system effectively extends the driver's range of perception as far as road conditions are concerned. In addition to directly helping the driver navigate through a busy road, it will also reduce the stress of driving, thus bringing in positive correlation into the overall experience. The net outcome of this assistance will be a reduced probability of accident.

The goal of our work reported in this paper was to establish the set of criteria for an effective implementation of a driving assistant, to define such a system, and to build its working model from inexpensive off-the-shelf components. The implementation was meant to yield further insights into the problem. For the sake of feasibility and easy demonstrability, the system has been installed on a ride-on toy car; however, its scalability to a "real" vehicle has not been lost from sight. One of our objectives was to make sure that the assistant could be in principle installed on any vehicle without permanent modifications to its exterior. For this reason, its sensor modules communicate over a wireless link and are powered of independent batteries. This way, they can be trivially attached to the vehicle's exterior, e.g., with magnets, without the need of wiring them to the data collection hub.

The remainder of the paper is organized as follows: Section (2) discusses the current technology in driving-aid devices. Section (3) emphasizes on our proposed solution and the implementation of major functional blocks. Section (4) describes system-level integration. Section (5) focuses on the overall test plan. Section (6) elaborates on system response time and possible sensing failure. Section (7) suggests follow-up work of our design. Finally, Section (8) summarizes the main contributions of our proposed solution.

2. Problem Statement

A. Existing Technology

Most of the academic research on driving support systems focuses on automating the actual driving rather than assisting the driver by enhancing its awareness of



the car's surroundings [7] [8 to 16]. None of such automated driving systems are currently available to the public. On the other hand, several types of practical driving-aid devices have been developed and made available on the market. Blind-spot detectors, lane-change assistants, and back-up/parking sensors are a few examples. The current technology in driver support systems is diverse in terms of functionality, methodology, and implementation. The common denominator of all those systems is obstacle detection by sensing: to this end they utilize various forms of sensors, such as laser light, ultrasonic, radar, infrared, and CCD cameras. For example, the parking support system from Backup-Sensor.com [17] places four ultrasonic sensors at the lower edge of the rear bumper. The blind-spot monitor from Valeo [18] deploys one multi-beam radar (MBR™) sensor on each side of the vehicle. Another blind-spot detector, from Xilinx [19], employs infrared sensors. The comprehensive and complex driving-aid system, put forward by the Lateral Safe project [20], integrates a long-range radar, an ensemble of short-range radars, and a set of stereovision cameras using sophisticated sensor fusion techniques.

The existing driving-aid devices employ various types of indicators, such as beeping sounds and LED lights, to issue warnings to the driver. The parking system designed by Backup-Sensor.com [17] comes equipped with piezo-speakers whose beeping frequency increases as the car approaches an object. The Valeo blind-spot detector [18] emits alarm sounds and flashes LED icons on the rear view mirrors.

B. Areas for Improvement

An overview of the currently available driving-aid devices reveals a certain trend. Simple and inexpensive systems tend to focus on one specific area of the vehicle (effectively targeting a certain type of maneuver), e.g., the blind spots or the rear bumper. On the other hand, the more comprehensive systems, purporting to monitor the entire perimeter of the vehicle, are extremely complicated and thus very expensive [21]. Their decisions (based on data fusion from diverse sensors) are driven by complex rules involving AI techniques, which make them inherently uncertain and error prone [22]. In other words, there are many points where such a system may fail. Consequently, it cannot be both reliable and inexpensive at the same time.

The alert signals issued by the existing systems are limited to the audio-visual range, i.e., those signals consist of various beeps and lights triggered by the detected hazards. These days drivers tend to be more and more exposed to this kind of input, as the various personal devices (e.g., cellular phones, iPods, PDAs) compete for our attention. One can argue that the most likely type of driver being in a desperate need of reliable driving assistance is a person addicted to such devices and thus routinely overwhelmed by their noise. Such people will be inclined to subliminally push that noise into the background and, if not simply ignore it then at least fail to consider it instinctively a "true" alert.

The visual domain is not much better. Even ignoring the contribution of visual input from satellite radios and GPS navigators (which also constitute a considerable source or

distracting audio input), the crowded metropolitan roads abound in numerous visual distractions. Advancements in technology will only tend to bring more of those, e.g., in the form of aggressive plasma displays, whose level of distraction is orders of magnitude above that of the old-fashioned marquee lights.

Consequently, we envision two areas for improvement. First, our goal is to build a simple and inexpensive general-purpose system capable of parking assistance, as well as providing support during normal driving. Second, we would like to provide an extra level of sensory input, namely tactile feedback, whose role would be to clearly separate the system's alerts from the background noise. Additionally, the wireless component of our system brings about the added value of flexibility, making it easy to install the driving assistant in any vehicle in a non-intrusive way (e.g., without drilling holes and/or modifying the vehicle's wiring). As a byproduct of our exercise, we shall validate the feasibility of a low-cost, low-power radio channel for replacing wires in a sensor-based driver aid system.

3. The Proposed System and its Implementation

As stated in the previous section, our main objective was to define and functionally verify a new inexpensive driving assistant addressing the need for broader area coverage and more effective stimulation for the driver. The design features a network of sensors mounted around the exterior body of the vehicle. The system also employs several types of actuators to indicate the different levels of hazard to the driver.

The proposed system contains five physically separate components: four of them are the sensor modules, called *nodes* in the sequel, to be attached to the car's exterior, and the fifth one is the *controller* responsible for coordinating the operation of the nodes, collecting data from them, and presenting alerts to the driver. The sensor modules cover the two front corners and the blind spots on both sides of the vehicle. While the blind spot areas are the most likely ones to be neglected by drivers, the two front corners are found to produce the highest accident rates [20]. All five modules are built around the same processing unit consisting of a microcontroller and a radio transceiver.

Figure 1 shows the correlation among the system components. The sensor modules employ ultrasonic sensors, while the controller connects to three types of indicators: the LEDs panel, the buzzer, and the vibrators mounted on the steering wheel.

A. The Sensors

The function of the sensor network is to detect objects appearing in the four zones around the vehicle. Those objects can be stationary (located on the side of the road), e.g., trees or buildings, or mobile, such as other vehicles or pedestrians.

The driving assistant detects objects in the monitored areas via ultrasonic proximity sensors. Infrared and radar sensors were considered as possible alternatives. However, infrared sensors lack the accuracy of their ultrasonic counterparts due to the ambient noise and



infrared radiation [23]. Radar sensors, on the other hand, are expensive and complicated in use, because of the need for CPU-intensive image processing techniques. In addition to raising the project cost directly (by being expensive themselves), they would also need a more complex processing platform, which would also increase the cost of their encasing sensor modules.

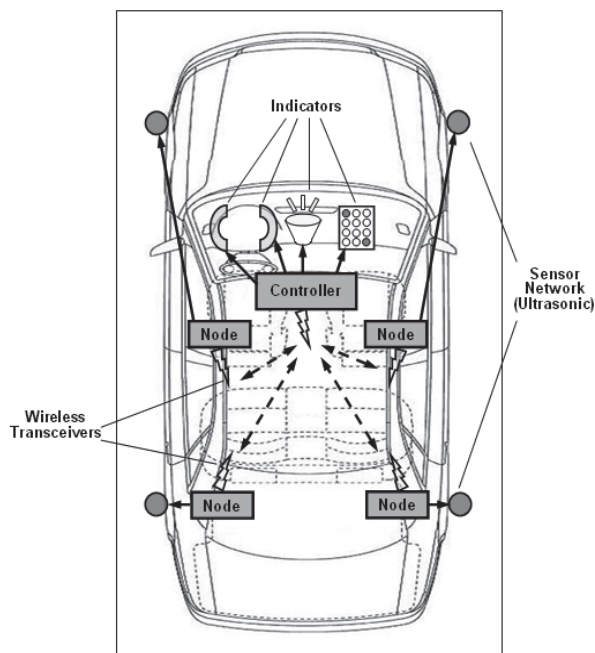


Figure 1. Layout of the driving assistant (adapted from: http://www.suurland.com/blueprints_archive.php)

Consequently, we based our design on ultrasonic sensors, specifically the PING)))TM sensors from Parallax [24]. Such a sensor is in fact a compound device consisting of an ultrasonic emitter and the “proper sensor” whose role is to detect the reflected signal sent by the emitter. The device is triggered by a pulse from the microcontroller forcing it to emit a 40 kHz burst to the environment. Following the end of the burst, the microcontroller is expected to measure the amount of time elapsing until the arrival of the reflected signal, as perceived by the sensor. Figure 2 shows a time diagram illustrating the typical operation of a PING))) sensor and defines some key parameters. The simple interface consists of a single pin operating in two directions. Initially, the pin acts as input to the sensor. A measurement starts when the microcontroller pulls the pin high for a short interval t_{TRIG} (the SIG phase) and then switches to monitoring the pin status. In response, the sensor will pull the pin low while it emits the ultrasonic burst, which takes the t_{HOLD} interval. Then, the sensor pulls the signal high (the t_{ECHO} phase) until it perceives the reflected signal, at which time it will pull the pin down. Thus, t_{ECHO} directly determines the distance between the sensor and the reflecting object. The amount of time separating the end of a measurement (the signal going down) from the beginning of the next one (the initial pulse sent by the microcontroller) is denoted by t_{DELAY} .

To study the effective detection range, a PING))) sensor was placed on a table, such that its backplane was

perpendicular to the table surface. Then the test object, a rectangular box with the surface dimensions of 14.7 cm by 22.0 cm, was positioned at a distance of 10 cm straight away from sensor. After taking 10 sample readings, the test object was moved 10 cm further away and another 10 readings were taken. The same procedure was repeated at 10 cm intervals until the test object was moved 230 cm away from the sensor. The surface of the test object remained parallel to the sensor backplane in all cases. The results plotted in Figure 3 show the correlation of the measured value of t_{ECHO} with the actual distance of the object from the sensor. The delay is expressed as the *count* value accumulated in an integer counter while waiting for the respective pin to become low.

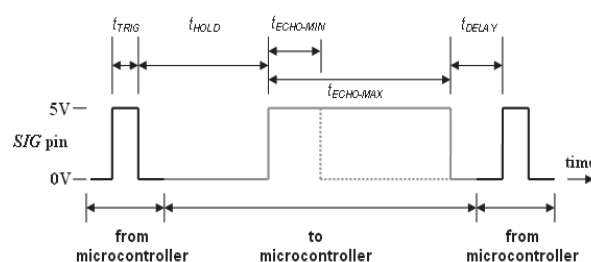


Figure 2. Operation time diagram of the PING))) ultrasonic sensor

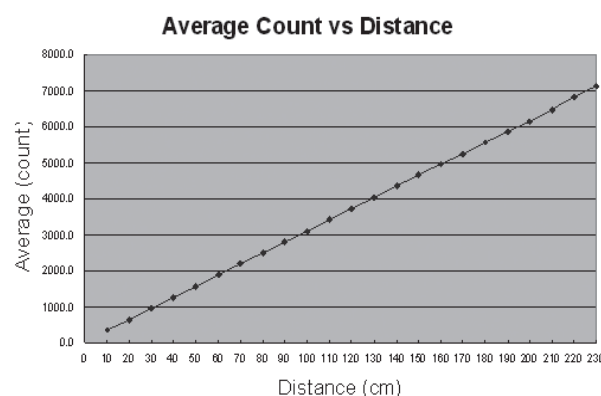


Figure 3. Correlation of t_{ECHO} with distance for the PING))) sensor

Although Figure 3 shows a perfectly linear relationship between the measured delay and the distance, our experiments indicated excessive noise when the distance exceeded 200 cm. In other words, even though the sensor is in principle capable of detecting objects that are over 200 cm away, the measurement accuracy degrades drastically as the distance increases over that range.

The second series of experiments was aimed at establishing the directional characteristics of the sensor, i.e., the dependence of its indications on the angle. We started by drawing on the table concentric half circles with the common center positioned at the sensor. Then, the test object, a pencil with 0.4 cm diameter, was moved along the circumference of the circle starting from angle 0° (straight away) to the point at which the sensor was unable to detect its presence. The procedure was repeated for half circles of the radii from 10 cm to 60 cm at 5 cm intervals. The results are plotted in Figure 4. For most distances less than 60 cm, the detectable angles vary between 10° and 20°.



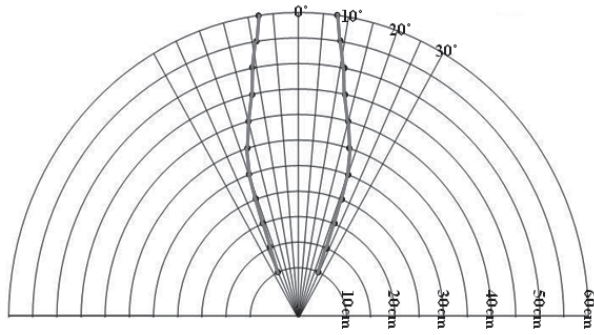


Figure 4. The directional characteristics of PING)))

TABLE I
ZONE DESIGNATION

Zone	Distance from car	Number of car widths	Danger level
0	> 60cm	> 1	safe
1	30cm – 60cm	0.5 - 1	mildly dangerous
2	< 30cm	< 0.5	very dangerous

Following our understanding of the performance of the ultrasonic sensors, the next step involved making design decision regarding the specific areas of sensor coverage. The ride-on toy car used as the prototype model has a width of approximately 60 cm. The area surrounding the car was partitioned into three zones listed in Table I. Each of the four targeted areas: the left blind spot, the right blind spot, the left front corner and the right front corner, was assigned one sensor module. Using the zone designations from Table I and the results from Figure 4, Figure 5 visualizes the areas of coverage for the entire sensor network. The sensor numbering from Figure 5 will be followed in the rest of the paper.

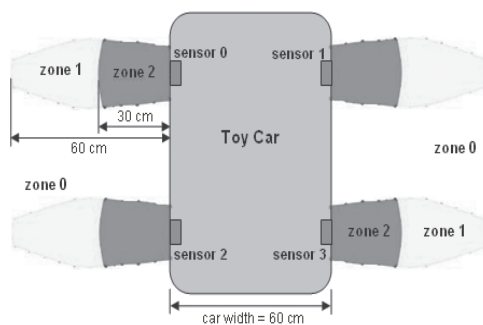


Figure 5. Coverage area for the ride-on toy car

B. The Indicators

The goal of the indicators is to warn the driver about the presence of a potential hazard. Three different indicators were implemented in the prototype system: an LED display panel, two tactile vibrators mounted on the steering wheel, and a buzzer. Different indicators are activated depending on the degree of hazard. All indicators connect to and receive signals from the controller.

The LED display is built of an off-the-shelf 5x8 LED matrix shown in Figure 6. Four LEDs were assigned to directly represent the status of the four ultrasonic sensors,

based on their position: they have been highlighted in Figure 6. Thus, LED 0 corresponds to the status of the left front corner sensor (sensor 0); LED 1 corresponds to the right front corner sensors (sensor 1), and so on.

The vibrator motors used in the project are manufactured by Jameco Electronics [25]. Their voltage/current requirements make it possible to drive them directly from the microcontroller, with no need for a power amplifier.

A vibrator may vibrate continuously or intermittently. In the second case, the vibrations are produced in 2-second spurts separated by 2-second breaks.

The audio signal is provided by the magnetic buzzer from CUI Inc. [26] operating as shown in Figure 7. To trigger the buzzer, a square wave of 2400 Hz frequency and 1-3 V amplitude should be fed to the circuit input. A reasonable approximation of such a waveform is generated by the microcontroller. Similar to the vibrator, the buzzer can generate continuous and intermittent sounds. In the latter case, the buzzer sounds five times, each time for a duration of approximately 0.5 s.

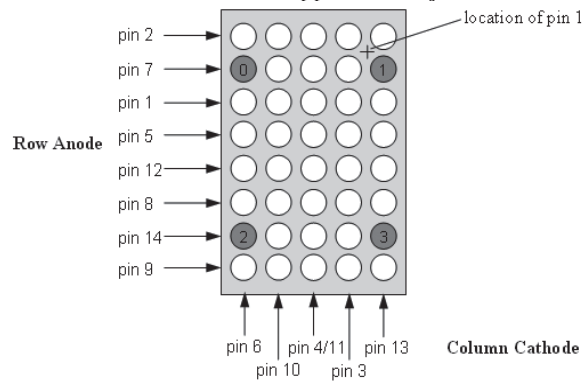


Figure 6. The LED panel

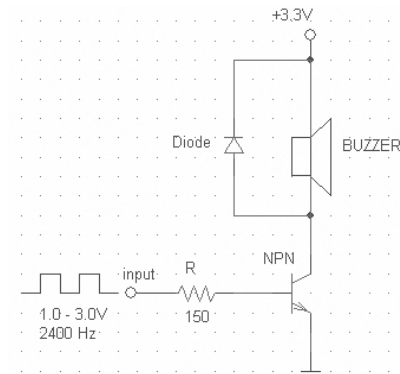


Figure 7. The buzzer circuit

Of the three types of indicators, the vibrators and the buzzer generate stronger and more direct stimuli than the LED display. The driver should immediately hear the buzz sound or feel the vibrations, as soon as a hazard is present. LED flashes can only be perceived if the driver looks directly at the display. Their role is to indicate the specific location of the hazard.

For as long as no obstacles are present in zone 1 and zone 2, the vehicle is considered safe, and none of the indicators are activated. When an object is present in zone 1, the vehicle is considered only “mildly threatened,” since the closest foreign object is half to one vehicle width away. At this time, the respective LED



indicators are turned on to provide a “low-level alert” signal to the driver.

When an object falls into zone 2, the vehicle is considered “threatened.” At this time, in addition to the LED indicators, either the vibrators or the buzzer are turned on, according to Table II. Multiple sets of indicators may be activated if multiple objects appear in the different danger zones. In addition, based on the user’s preference, two types of operation modes, *continuous* and *intermittent*, can be selected for activating the vibrators and the buzzer.

TABLE II
RESPONSES OF INDICATORS IN VARIOUS SCENARIOS

Sensor Number	Zone	Activated Indicator
0	0	None
	1	LED 0
	2	LED 0, left vibrator
1	0	None
	1	LED 1
	2	LED 1, right vibrator
2	0	None
	1	LED 2
	2	LED 2, buzzer
3	0	None
	1	LED 3
	2	LED 3, buzzer

C. Wireless Communication

Five wireless transceivers are present in the prototype system: one on each of the four sensor nodes and one on the main controller. In principle, one-way communication would be sufficient, with the sensor nodes periodically notifying the controller about their status. However, with this simple scheme, every sensor node would have to operate constantly, sending its measurements at a high rate, because 1) it would have no choice but to assume that its reports are always needed, 2) its only assurance of success (with no feedback from the controller) would be to send the same data over and over again. Such a mode of operation would be congestion-prone and poorly scalable to a larger population of sensors. Moreover, batteries at the sensor nodes would deplete quickly, which would turn the advantage of the wireless system into a nuisance.

While the prototype system itself is not particularly frugal with respect to energy consumption, its organization facilitates power efficient operation of the prospective next (more practical) version. To do it properly in the prototype, we would have to build custom boards, e.g., better integrating the radio module with the microcontroller, which was not considered immediately relevant. Both the microcontroller (MSP430F1611), as well as the radio module (CC1100), are capable of highly power efficient duty cycling, i.e., reducing energy usage to a negligible level while maintaining responsiveness to requests from the controller.

The first prerequisite to this kind of operation is two-way communication enabling the controller to convey requests to the nodes. In particular, if the driver decides that the current driving conditions require no assistance, the controller may switch the network to a power saving mode. It may also selectively disable some of the nodes, e.g., front, rear, left, right, depending on the circumstances.

With a reasonably general approach to communication between the sensor network and the controller, the system is open for extensions. For example, a wireless temperature sensor can be added to the system and the controller appropriately augmented to display its readings. The CC1100 transceiver module operates within the 868 MHz frequency band with 256 selectable channels. A complete reference design that can be directly interfaced to a microcontroller (as a sister board) is available from Texas Instruments as CC1100 EM (evaluation module). The MSP430F1611 microcontroller is programmed under PicOS [27], which is a compact real-time operating system with powerful networking capabilities [28]. PicOS comes equipped with a driver for CC1100; thus, our implementation of wireless communication in the driving assistant was straightforward and mostly consisted in deciding on the collection and format of the messages to be exchanged between the controller and the nodes.

Essentially, there are three types of such messages: control, data, and acknowledgment – see Figure 8. Control messages are sent from the controller to a sensor node, data messages travel in the opposite direction; finally, an acknowledgment confirms the receipt of a message. The simple protocol assumes that the sender of a data or control message retransmits the message at randomized short intervals until it receives an acknowledgment.

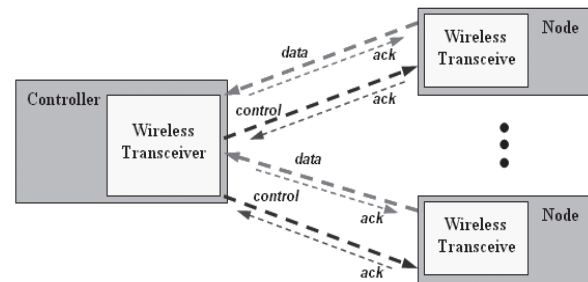


Figure 8. Information flow in the wireless network

Each message is sent in a single packet whose generic format is shown in Figure 9. The GID (group Id) field identifies the system, i.e., one particular network. Multiple networks using different GID values can safely coexist within their mutual range, even if they share the same RF channel. The transmission power of CC1100 can be adjusted as to reduce the likelihood of interference with a device outside the vehicle perimeter, while still providing for reliable communication within the system. Notably, PicOS offers inexpensive packet encryption facilitating cryptographically secure separation of networks. This option has not been used in the present project, but is available as an easy add-on to a future version of the driving assistant.

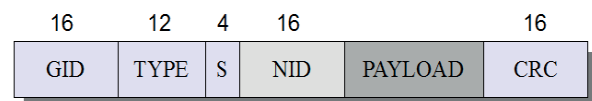


Figure 9. Packet format



Every packet ends with a 16-bit CRC code calculated according to the ISO 3309 standard. Generally, the interpretation of its remaining components depends on the TYPE field. Intentionally, S provides a short (modulo 16) serial number field to be used for detecting missed packets in a sequence and/or pairing acknowledgments with their packets. TYPE zero identifies acknowledgments.

All packets in our present application use the NID field to indicate the node number to which the packet pertains. For a control packet, this field contains the number of the sensor node to which the packet is addressed. For a data packet, the NID field identifies the sender (the packet is implicitly addressed to the controller). For an acknowledgment, NID relates to the sensor node involved. That is, if the acknowledgment is sent by a node, NID identifies the sender, if it is sent by the controller, then NID refers to the node to which the acknowledgment is addressed.

The payload field of a control packet (TYPE = 1) contains two 16-bit values: t_{DELAY} (see Figure 2) and the number of readings per sample n_r . Following the reception of such a packet, the sensor node will begin periodic measurements separated by t_{DELAY} milliseconds. After every n_r distance measurements, the node will use their average to assess the hazard according to Table I. The node can be switched off by requesting $n_r = 0$.

The payload of a data packet (TYPE = 2) carries the zone report, i.e., 0, 1, or 2, encoded as a 16-bit integer. To avoid unnecessary congestion, such a packet is only sent if the report would be different from the previous one, i.e., something has changed. Note that data packets are acknowledged: they are sent persistently until the node makes sure that the controller has been informed. Thus, there is no need to notify the controller unless the situation has changed with respect to the previous successfully conveyed report. Should the controller wish to verify whether a particular node is operational, it can do so with a control packet, which again will be acknowledged by the node.

Acknowledgement packets (TYPE = 0) have no payload (the PAYLOAD field is empty). They are matched to the packets being acknowledged by the S field. A node sending a data packet is expected to increment that field by 1 (modulo 16) with every new report. Similarly, the controller is expected to sequence its control packets in the same way, on the per-node basis.

D. The Modules

Two types of modules are present in the system: the nodes and the controller. The controller receives data packets from the nodes, interprets their alerts, and activates the proper response of the indicators. The controller also receives direct commands from the driver through a series of DIP switches integrated onto the microcontroller board. The node modules receive control packets from the controller, interpret its commands, carry out their measurements, and report situation changes to the controller. Figure 10 depicts the data flow within the two types of modules.

All the modules of our prototype system have been built of five identical MSP430F1611 development boards. They have been programmed under PicOS [27], and they

run multi-threaded reactive applications (called *praxes* in PicOS) to ensure that all commands are processed and all responses are generated in real-time.

A status LED is present on each node board to indicate whether the sensor is currently running or has been stopped. Even though the ultrasonic sensors come with their own LED indicators, those indicators are obscured from the driver's view by the mounting. The status LEDs were useful for tests and debugging. A real-life version of our system will need no such feature, as the status of a sensor module can be reliably determined from the controller.

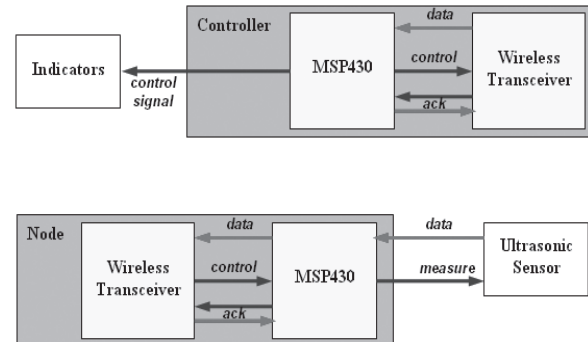


Figure 10. Flow of data in the two types of modules

A set of DIP switches added to the controller module offers a simple user interface to the system. For example, the driver may want to disable some or all the sensors, depending on the current driving conditions. It is also possible to switch off some of the indicators, e.g., the buzzer, or select the vibrator mode (continuous versus intermittent).

4. System Integration

The prototype system was installed on a ride-on toy car for demonstration and validation. The dimensions of the car are roughly 85 cm long and 60 cm wide. The controller was placed on the dashboard, with the vibrators mounted on the steering wheel. For the ease of mounting and adjustments, the sensors were separated from their modules and connected with wires: the modules were placed inside the car, while the sensors were attached to the exterior at the height of about 35 cm from the ground. The fully equipped car is shown in Figure 11. Table III provides the complete list of parts.

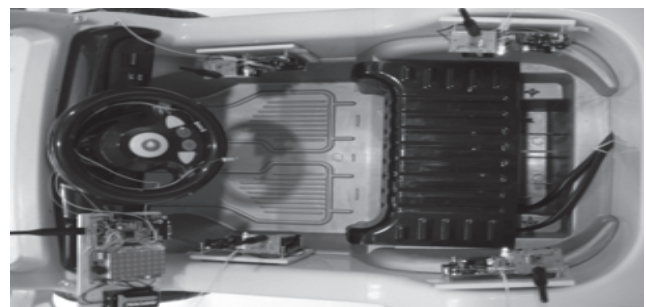


Figure 11. The prototype car



TABLE III
COMPLETE PART LIST

Component	Manufacturing #	Quant.
Microcontroller Development Board	MSP430F1611	5
PING))) TM Ultrasonic Sensor	Parallax Inc.	4
Ride-on Toy Car		1
LED Matrix (5X8, Red)		1
Buzzer (2.4 kHz)	CEM-1201S	1
Diode		1
NPN Transistor	2N3904	1
Vibrator Motor (3V, 75mA, 8500 rpm)	6ZK053	2
Wireless Module	CC1100EMK868-915	5
Antenna	CC1100EMK868-915	5
Connectors	852-10-100-10-001000	2
Battery Holder (9V, Wire Leads)		5
Alkaline Battery (9V)		5
Voltage Regulator (+5V, 1A)	7805UC	4
Voltage Regulator (3.3V)	LT1086CT	5
Tantalum Capacitor (10 μ F, 10V)		15
Ceramic Capacitors (0.33 μ F, 50V)	C320C334M5U5TA	4
Ceramic Capacitors (0.47 μ F, 50V)	C322C474M5U5TA	4
LED (Green, Clear)	LTL-4238	4
DIP Switch (Rocker, Sealed, 2 Positions)	76SB02ST	1
DIP Switch (Rocker, Sealed, 4 Positions)	76sB04ST	5
Resistor (150 Ω)		1
Resistor (4.7 k Ω)		4
Wood Blocks		5

A. Power Circuits

During the development stage, all the components, circuits and microcontrollers were powered from adaptors connected to the power outlet. To allow the system to be installed on the toy car and freely moved along with it, portable power supplies were built.

For each node, two output voltage levels, 3.3 V and 5V, are generated in the power circuit. While 3.3 V is used to power the MSP430F1611 microcontroller and most components on the node, 5V is needed by the PING))) ultrasonic sensor. As a result, the power circuit consists of a 5V and a 3.3 V voltage regulator connected in parallel. One has to notice that this is a rather serious disadvantage from the viewpoint of power-efficient operation of the wireless sensor node, as each regulator incurs unavoidable losses. However, there exist ultrasonic sensors operating at 3.3 V, e.g., the LV-MaxSonarTM range finder from MaxBotic. Notably, the range of that sensor is over 6 m, which makes it better suited for deployment in a realistic vehicle. With the unification of voltage requirements, the regulators can be completely eliminated. This is possible because the microcontroller, the RF module, as well as the low-voltage ultrasonic sensor exhibit comfortable tolerance regarding the range of supply voltage, with the lower bound around 2.5V.

Figure 12 shows the power circuit for the sensor node. The bypass capacitor C3 filters out the addition noise. In

the controller, the power circuit is simpler, since only a single 3.3 V voltage level is required to power the microcontroller and all the components.

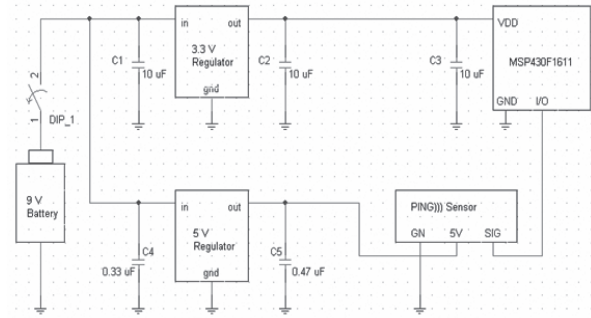


Figure 12. The power supply circuit for the node

B. The Software

The software for the controller and the nodes consists of two PicOS applications (praxes) organized into multithreaded programs. In the controller, the *root* process (which in a PicOS praxis plays the role of “main program”) coordinates the activities of three child processes: *sender*, *receiver*, and *actuator*. The *sender* process is only created if a control packet has to be sent out and remains active until the last outstanding control packet has been acknowledged by the recipient. The process runs automatically on startup to initialize the sensor nodes; then it is invoked by *actuator* whenever the parameters of a sensor node should change.

The *receiver* process receives data and acknowledgement packets from the nodes and sends out acknowledgement packets to the nodes to confirm the receipt of data packets. The process operates in an infinite loop and continuously expects to receive incoming packets.

The *actuator* process controls the operations of the three indicators: the LED display, the vibrators and the buzzer. The process also continuously monitors the status of the DIP switches. Data packets received by *receiver* update the global variables to indicate the most recent status of the sensors. The process operates in an infinite loop, reading the global variables in every iteration and updating the indicators accordingly to match the current status of the sensors.

All nodes run exactly the same program, which differs by a single constant at every node: the node Id (0 – 3, see Figure 5). Having initialized the module, the *root* process creates one child process, *receiver*, which remains active for as long as the node is powered up. The *receiver* process receives control packets and acknowledgements from the controller. It also sends out acknowledgment packets to the controller to confirm the receipt of control packets. If, based on the controller’s request, the node is to be active, i.e., it has to carry out its sensing duties, *receiver* creates an additional process, named *sensor*, whose role is to perform the measurements and report zone alerts to the controller. During its lifetime, the *sensor* process continuously triggers the ultrasonic sensor and measures the value of t_{ECHO} (see Figure 2). If the new sample shows that the zone of the closest foreign object has changed, the *sensor* process creates and sends out a data packet addressed to the controller.



5. Test Plan

A test plan was designed to ensure that the prototype functioned towards our expectation. Many test cases have been considered to verify all functionality aspects of the system.

A. Test Cases

The test cases considered in the test plan can be classified into several categories. A single-object test focuses on only one foreign object appearing in the vicinity of the vehicle. Multiple-object tests deal with two or more objects appearing simultaneously. Object type tests deal with stationary and mobile objects. Mode tests verify the *continuous* and *periodic* modes of operations for the vibrators and the buzzer.

Finally, a test drive of the prototype system was also conducted to get a feel of how the system performs while moving in its environment. Table IV summarizes the test cases considered in the test plan.

TABLE IV
LIST OF TEST CASES

Category	Description
Single Object	object in zone 0, 1 and 2 of sensor 1
	object in zone 0, 1 and 2 of sensor 2
	object in zone 0, 1 and 2 of sensor 3
	object in zone 0, 1 and 2 of sensor 4
	objects that trigger more than 1 LED
Multiple Objects	objects that trigger an LED plus the buzzer
	objects that trigger an LED plus one vibrator
	objects that trigger both vibrators
	two objects that both trigger the buzzer
	stationary object: a wall
Object Type	moving object that passes through zones 2, 1, 0 of a single sensor
	moving object that passes through scanning areas of different sensors
On/Off-Switch	turn sensors on and off
	turn vibrators on and off
	turn buzzer on and off
Mode	vibrator and buzzer operations in <i>continuous</i> mode
	vibrator and buzzer operations in <i>periodic</i> mode
Test Drive	moving the car around in the environment

B. Test Settings

The prototype system has been tested in the open space near the IRMACS Center reception in the Applied Sciences Building of Simon Fraser University. The experiment was set up such that, except for the objects under test, no other obstacles came within 1 m on both sides of the car. Cardboard boxes of various sizes were used as test objects, and video clips were taken to demonstrate the various functionality aspects of the system.

6. Discussion

A. Response Time

With t_{DELAY} set to 0.125 s, each ultrasonic sensor takes a measurement approximately once every 0.125 s. Only one reading is taken during each sample.

Ultrasonic waves travel through the air at approximately 1130 feet per second or, equivalently, 1 cm in 29.034 μ s. When an object is present at the boundary of zone 0 and zone 1 (equivalent to 60 cm away from car), the

approximate time for the wave to travel from the sensor to the object and back is:

$$t_{ECHO} = \frac{29.034 \mu s}{1 cm} \times 60 cm \times 2 = 3.484 ms. \quad (1)$$

Adding the trigger pulse duration, hold time, and delay time (see Figure 2), the total sampling time required to detect an object at 60 cm is:

$$\begin{aligned} t_{SAMPLE} &= t_{TRIG} + t_{HOLD} + t_{ECHO} + t_{DELAY} \\ &= 8 \mu s + 750 \mu s + 3.484 ms + 0.125 s \\ &= 0.129242 s \\ &= 0.129 s. \end{aligned} \quad (2)$$

Figure 13 shows the observed system delay, t_{SYS} , needed for a pulse generated by a node controller to reach the indicators on the controller side via the wireless transceivers. CH 1 is connected to an I/O pin on the node, and CH 2 is connected to a vibrator control pin on the controller. The system delay includes the combined delay in the microcontrollers and the wireless network. It can be estimated by computing the average of the delays shown in Figure 13, i.e.,

$$t_{SYS} = \frac{t_{RISE} + t_{FALL}}{2} = \frac{80 ms + 240 ms}{2} = 160 ms. \quad (3)$$

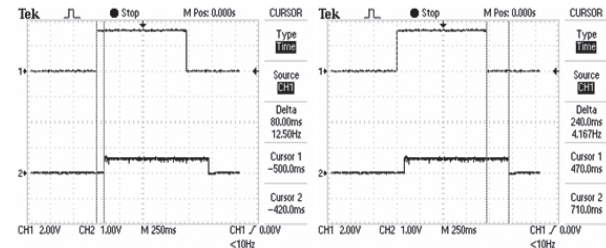


Figure 13. System delay

Adding (2) and (3), the total system response time is found to be:

$$t_{TOTAL} = t_{SAMPLE} + t_{SYS} = 0.129 s + 0.160 s = 0.289 s. \quad (4)$$

In summary, the maximum system response time for objects appearing in the danger zones is 0.289 s. In other words, the system is expected to activate the appropriate indicators within 0.289 s after an object appears in any of the danger zones.

To assess the adequacy of this delay, let us carry out a simple analysis. Suppose an object with velocity V_O and acceleration a_O is approaching the car d_{HORIZ} cm away, at a contact angle of θ_i , as illustrated in Figure 14. The car is traveling at velocity V_C . By trigonometric relations:

$$d = \frac{d_{HORIZ}}{\sin \theta_i}, \quad (5)$$

$$V_C = \frac{V_O}{\cos \theta_i}. \quad (6)$$



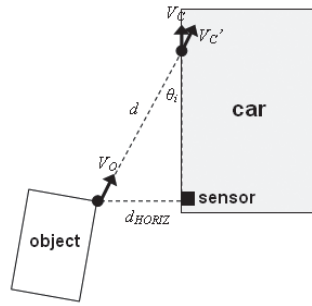


Figure 14. Geometric relation between the car and an approaching object

The time t until the object hits the car is determined using the kinematics relation:

$$d + V_C t = V_O t + \frac{1}{2} a_O t^2. \quad (7)$$

Thus, t is also the amount of time available for the system to react to the approaching object. Rearranging the terms and solving for time, we get:

$$t = \frac{(V_C' - V_O) \pm \sqrt{(V_O - V_C')^2 + 2a_O d}}{a_O}. \quad (8)$$

Since the negative term is not applicable, (8) boils down to:

$$t = \frac{(V_C' - V_O) + \sqrt{(V_O - V_C')^2 + 2a_O d}}{a_O}. \quad (9)$$

By plugging various sets of V_O , V_C , a_O , θ_i , and d_{HORIZ} into (5), (6) and (9) we can obtain the respective t . For simplicity, d_{HORIZ} is fixed at 60 cm because the prototype system considers any distance less than 60 cm as dangerous and starts issuing warnings to the driver. Also, the car is assumed to travel at the typical speed of 60 km/h.

The actual system response time given by (4) is compared against the allowable system response time obtained from (5), (6), and (9). The actual system response time is found to be shorter in most cases. Therefore, the indicators are triggered before the approaching object strikes the vehicle. Table V shows the maximum object velocity the system is capable of handling at various contact angles and under various accelerations.

Table V can be interpreted as follows. For an object traveling at a contact angle of 2.5°, 5.0° and 7.5° and an acceleration of 4.0 m/s² or less, the indicators are activated ahead of collision if the object velocity is less than 110 km/h. For an object traveling at a contact angle of 10.0° and an acceleration of 4.0 m/s² or less, the indicators are activated ahead of collision if the object velocity is less than 100 km/h. For an object traveling at a contact angle of 20.0° and an acceleration of 4.0 m/s², the indicators are activated ahead of collision if the object velocity is less than 80 km/h.

Table V shows that the system is able to react to approaching objects in a timely manner, even though the

cases considered in the table are rather extreme. Since objects traveling at a large contact angle (more than 10°) and high acceleration (over 3.0 m/s²) are extremely rare in real traffic, the system performance is deemed adequate for handling the majority of hazardous situations on the road.

TABLE V
MAXIMUM ALLOWABLE OBJECT VELOCITY

Maximum Object Velocity (km/hr)	Contact Angle (deg)					
	2.5	5.0	7.5	10.0	15.0	20.0
0.00	110	110	110	90	90	85
1	+	+	+	90	90	85
0.5	110	110	110	90	90	85
+	+	+	+			
1.0	110	110	110	90	90	85
+	+	+	+			
1.5	110	110	110	90	90	80
+	+	+	+			
2.0	110	110	110	85	85	80
+	+	+	+			
2.5	110	110	110	85	85	80
+	+	+	+			
3.0	110	110	110	85	85	80
+	+	+	+			
3.5	110	110	110	85	85	80
+	+	+	+			
4.0	110	110	110	85	85	80
+	+	+	+			

B. Failures

Occasionally, the indicators may fail to provide the correct representation of the area around the vehicle. An indicator may remain deactivated when it should have been activated, or it may be activated when it should remain silent. The former case is usually more serious because it puts the vehicle at risk.

Several problems may affect the operation of the driving assistant. For example, packet congestion in the wireless channel may delay the receipt of a data packet. Other ultrasonic waveforms and noise signals present in the environment may affect sensor readings. The type of surface of the intruding object may also influence the reflectivity of ultrasonic waves.

No driving aid system can be 100% foolproof. Considering the possibility of failure, the system should only be used as an *assistant* rather than a substitute for staying alert, maintaining safety distance, performing mirror and shoulder checks, etc.

7. Future Work

As the first design of the driving assistant, the prototype system has room for improvement in several areas.

A. Sensor Network

The number of ultrasonic sensors in the sensor network can be increased to provide better coverage around the vehicle. In particular, the rear end and the front of the car require some form of scanning in addition to the blind spots and the two front corners covered by the current system. Furthermore, numerical analysis techniques may be applied to filter out erroneous readings and reduce the effect of noise in the sensor measurements. Also, instead of using off-the-shelf components, device-specific



sensors can be designed and manufactured for use in the driving assistant.

Our experiments have indicated that the wireless network is a reliable medium for the type of system represented by our driving assistant. In particular, we have never experienced congestion problems resulting in prolonged packet loss that would affect the real-time status of an alert. This is because the alert notifications (in their present form) pose little demand for bandwidth. However, the situation may change when more sensors are added to the network, especially if the data sent by those sensors are more complicated than sequences of simple numbers. Note that any numerical processing should be done in the controller rather than the node. This is because 1) the controller is less power-constrained, 2) the processing capabilities can be centralized in one place, which will reduce the cost of the whole system. This approach, however, may increase the bandwidth required to convey the raw data to the controller and bring about congestion. Notably, the present communication protocol is open for natural improvements. First of all, considering the short distance of communication, the transmission rate can be increased (up to about 100 kbps for CC1100). Second, a TDMA scheme can be employed to guarantee bandwidth for critical nodes. The proper design clearly depends on the content and frequency of the requisite messages, but it appears feasible under all conceivable circumstances. Another advantage of the wireless sensor network is its futuristic provision for communication among different vehicles. By announcing and possibly synchronizing their maneuvers, multiple vehicles sharing the road may be able to arrive at a better social behavior, which would clearly translate into a reduced likelihood of collisions.

B. Indicators

A more elegant way of implementing the vibrators on the steering wheel would be to design a special fabric that produces tactile stimuli. The entire steering wheel can be wrapped with such a fabric, so the drivers may grip on any portion of the steering wheel, using one hand or both hands, and still feel the stimuli.

8. Conclusion

Starting from conceptualizing, designing, building to finally presenting, we have designed a driving assistant system which could be a possible solution for accident prevention. Three innovative features are characteristic of the prototype design discussed in this paper. First, an ultrasonic sensor system is implemented to detect foreign objects in areas around the blind spots and the front corners of the vehicle. Our experiments show that the low-cost PING)))TM ultrasonic yield satisfying accuracy in distance measurement. Secondly, an LED display, tactile vibrators, and a buzzer are introduced as the enhanced indicator design, which provides different levels of hazard warnings to the vehicle driver. Finally, a proprietary operating system, PicOS, and wireless interface, from Olsonet Communications Corporation, are evaluated and adapted in this project.

The final prototype design meets all the proposed functionalities of the new driving assistant. In addition, the control switches that are added to the main controller

further make the design user-friendly. The underlying software is structured so that many more sensor nodes can be easily integrated into the system. Finally, the work of this project can be further expanded to evaluate the traffic pattern of multiple vehicles, which may potentially become a valuable asset for future studies in traffic control and accident prevention.

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Implementation of ADALINE Algorithm on a FPGA for Computation of Total Harmonic Distortion of load current

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Abstract

The objective of the paper is to implement an ADALINE algorithm, for computation of Total Harmonic Distortion (THD) of harmonics in load current waveform on a Field Programmable Gate Array (FPGA). More power electronic converters (like SMPS, UPS, etc.) are added in the electrical distribution, due to their higher efficiency. But these converters, which are nonlinear loads, generate harmonics. Harmonics cause many ill effects in the power system and reduce Power Quality. Measurement of harmonics and computation of THD are important to assess the Power Quality. Harmonics have been measured by Fast Fourier Transform (FFT) executed on a microprocessor or a Digital Signal Processor. The FFT algorithm requires a minimum of 1000 samples from 10 cycles for a supply frequency of 50Hz. Neural Network algorithms are proving their efficacy in many complex situations. FPGAs are particularly suitable for implementing NN algorithms due to their hardware parallelism. This paper presents the results of implementation of ADALINE on an FPGA for computation of THD. The proposed implementation requires only 100 samples from 1 cycle of the load current. The results obtained using the proposed method are compared with that of the conventional FFT results and found to be very close.

Keywords: Harmonics, ADALINE, FPGA implementation.

1. Introduction

Harmonics levels are increasing in the power systems due to proliferation of non linear loads like static converters, saturated transformers, arc furnaces, fluorescent lights, etc.[1]. When the nonlinear loads are connected to the sinusoidal supply voltage, they draw current in non-sinusoidal manner [2]. A non sinusoidal load current is resolved using Fourier series into a sum

of many sinusoids of one fundamental frequency and other frequencies that are integral multiples of the fundamental [3] called harmonics.¹

The current harmonics produce more heating, additional losses and cause distortion in supply voltage. This has reduced the quality of power [4]. Examples of nonlinear loads are Variable speed drives (VSD), Uninterruptible power supplies (UPS) and SMPS, Arc and Induction furnaces, Fluorescent lamps with electronic ballasts, saturated transformers, etc. Measurement of harmonics is important for assessing the quality of power and also for subsequent filtering. The ill effects of harmonics are listed in section 2.

Different algorithms are used for measuring the harmonics. The Fast Fourier Transform (FFT) developed by Cooley and Tukey [5] laid the way for Fourier Transform to be executed on digital computers and subsequently on microprocessors. Other algorithms include spectral observer [6], Recursive DFT [7], Hartley transform [8], fixed gain Kalman filter [9] and variable gain Kalman filter [10]. A new class of algorithms based on the learning principles of neural networks evolved over a period, which includes the methods by Hartana and Richards [11], Mori *et al.* [12], Osowski S. [13], a linear adaptive neural network based on ADALINE by Dash P.K. *et al.* [14] and a Multi Output ADALINE [15].

This paper discusses the implementation of ADALINE algorithm using an FPGA. The current drawn by the SMPS of a personal computer, was sampled and acquired in real time using an ADC. 100 samples from 1 cycle were taken and fed to the ADALINE algorithm running on the FPGA kit. The results of harmonics values obtained using this method are compared with the results obtained for the same input signal by a

¹ This work was carried out VLSI Lab of Kongu Engineering College.



standard Power Quality Analyser Model C.A 8332 made by Chauvin-Arnoux, France [16].

Section 2 lists the ill effects of harmonics. Section 3 explains the International Electrotechnical Commission (IEC) standard, IEC 61000-3-2 standard for harmonic limits. Section 4 explains the resolution of a non sinusoidal wave into fundamental and harmonics. Section 5 briefs about the measurement of harmonics using ADALINE network. Section 6 is about the real time computation of THD using FPGA. Section 7 analyses the results obtained and Section 8 draws the conclusion.

2. The ill effects of harmonics

Harmonics, in supply voltage and load current, produce lot of detrimental effects, in the entire power system, spanning from generation through transmission to distribution.

The ill effects of harmonics are listed below:

- High current and hence more heating in neutral of 3P-4W system
- Overloading of power distribution transformers
- Reduced power factor and ampacity of conductors
- Failure of power factor correction capacitors
- Distorted Voltage waveforms to other loads
- Inductive interference into telephone lines
- Misfiring of AC and DC Drives,
- Nuisance breaker-tripping
- Incorrect meter functioning
- Failure of Protection relays
- Increased system losses as heat, etc.

Considering these ill effects, Utilities all over the world have specified the maximum limits for harmonics both in the system and in the loads. These limits are either directly or derived from two predominant standards, namely IEEE: 519 and IEC 61000-3-2. Utilities have also started imposing penalty for harmonics dumping by the users into the supply lines. Hence it is essential to identify the methods for measuring the harmonics and the standards specifying the limits for harmonics.

3. IEC standard for harmonic emission by equipment

The Institute of Electrical and Electronics Engineers IEEE: 519 standard [17] focuses on system level harmonic limits, which is in the purview of Utilities and electrical networks, in most cases. The IEC 61000 series standard specifies the limits for harmonic emission by equipment connected to the mains. The recommendations of these are applicable to electrical and electronic equipment with equipment input current in two ranges:

IEC 61000-3-2 standard [18] is for equipment with a rated current $\leq 16A$ and IEC 61000-3-12 standard [19] is for equipment with a rated current $> 16A$ but $< 75A$

per phase and intended to be connected to public low-voltage ac distribution systems of the following types:

nominal voltage up to 240V, single-phase, 2 / 3 wires;
nominal voltage up to 600V, three-phase, 3 / 4 wires;
nominal frequency 50 Hz or 60 Hz.

The equipments are classified into four classes:

Class A: Balanced 3-phase equipment and all other equipment except those in one of the following classes.

Class B: Portable tools

Class C: Lighting equipment including dimming devices with input active power above 25W.

Class D: Equipment having an input current with “special wave shape” and a fundamental input active power between 50 and 600W. This is the “high crest factor” waveform; e.g. single-phase rectifier input current waveform. Only TVs, PCs and PC monitors are included in this class.

The limits of harmonics for each phase of the line current are shown in Table 1. The limits are absolute. They are not related to power ratings of equipment. The limits are applicable to steady state harmonic currents.

Table 1. Limits specified in IEC 61000-3-2

Harmonic order (n)	Maximum permissible harmonic current (A)				
	Class A (A)	Class B (A)	Class C (% of I_0)	Class D	
				$75 < P < 600$ W (mA / W)	$P > 600$ W (A)
3	2.30	3.45	$30 \times \text{pf}$	3.4	2.30
5	1.14	1.71	10	1.9	1.14
7	0.77	1.155	7	1.0	0.77
9	0.40	0.60	5	0.5	0.40
11	0.33	0.495	11-39:3	0.35	0.33
13	0.21	0.315		0.296	0.21
$15 \leq n \leq 39$	$\frac{2.25}{n}$	$\frac{3.375}{n}$		$\frac{3.85}{n}$	$\frac{2.25}{n}$
2	1.08	1.62		Not specified	
4	0.43	0.645			
6	0.30	0.45			
$8 \leq n \leq 40$	$\frac{1.84}{n}$	$\frac{2.76}{n}$			



4. Measurement of Harmonics Using FFT Algorithm

Harmonics have been traditionally measured, either by RC tuned analog circuits, which do not satisfy the accuracy requirements, or by microprocessor based analysers employing FFT algorithm, which require more time. IEC:61000-4-7 [20] standard specifies the methodology for measuring harmonics; this standard uses the DFT algorithm which requires 1000 samples be taken over 10 / 12 periods of the 50 / 60 Hz power cycles and then computation be performed.

4.1 Time required for sample acquisition

The old IEC: 61000-4-7 Ed. 1.0 standard specified the methodology for measuring harmonics with the signal sampled for 16 complete periods of the voltage or current cycles. The revised IEC: 61000-4-7 Ed. 2.0 standard specifies that signal samples be acquired for 10 / 12 complete periods for 50 / 60 Hz systems instead of 16 periods. It further states that, a 1.5s smoothing filter be applied for all observation period. The response of the smoothing filter is given by,

$$S(t) = E(1 - e^{(-t/\tau)}) \quad (1)$$

For the 1.5s smoothing filter to be effective, a minimal observation period of 9.975s is needed. The arithmetic average of the measured values from the FFT time windows is calculated over the entire observation period.

4.2 Synchronisation needed for input samples

It is also necessary to accurately find the frequency of the input voltage or current waveforms to obtain the required accuracy in the computed values. The document 61000-4-7 requires a synchronous sampling of voltage or current signal, in order to limit leakage error and to assure reproducible results in presence of non-stationary signals. It is well known the output spectrum, calculated by means of FFT, is equal to the true one, only if the analyzed signal segment is stationary and if it is an integer multiple of its period [6]. The first requirement implies the non-variance of signal spectral content during sampling operation. The second requirement implies synchronous sampling. This means a sampling frequency which is a multiple of the fundamental signal frequency is to be used. When this condition is not satisfied, spectral leakage becomes visible. This error can be reduced performing a suitable windowing operation and/or using correction interpolation algorithms. However, these methods reduce but do not remove errors and need an extra computational cost; moreover, these procedures are accurate only when no harmonic and *inter-harmonic* signal pollution occurs. The IEC 61000-4-7 gives specific requirements on sampling synchronization. It requires a synchronous sampling of voltage or current signal, with a maximum permissible error of $\pm 0.03\%$ on the fundamental frequency. This requirement must be fulfilled within a range of at least $\pm 5\%$ of the nominal system frequency. Therefore an accurate

estimation of the fundamental frequency is required, even in presence of disturbances. At the same time, the standard does not suggest any measurement method for the instruments synchronization.

To have a close monitoring of harmonics in the system, accurate and fast measurement methods are preferred by both the utilities and users. Fast measurement of harmonics is also a mandate in Shunt Active Filters, intended to compensate for the harmonics.

This paper proposes a harmonics measurement system based on a class of Artificial Neural Network called ADALINE model. The method requires sample data from only 1 cycle of load current waveform for calculating the harmonics.

5. Measurement of Harmonics Using Adaline Network

In general, a nonsinusoidal signal can be expressed into a sum of sinusoids as,

$$y(t) = \sum_{n=1}^{\infty} A_n \sin(n\omega t + \phi_n) \quad (2)$$

This can be further expanded as,

$$y(t) = y_0 + A_1 \cos \omega t + A_2 \cos 2\omega t + \dots + A_n \cos n\omega t + B_1 \sin \omega t + B_2 \sin 2\omega t + \dots + B_n \sin n\omega t \quad (3)$$

y_0 is called the continuous (DC) component of signal $y(t)$; A_n and B_n are coefficients which represent the amplitudes of the n^{th} harmonics of the signal $y(t)$.

For signals which are *odd function* with *half wave symmetry* and *without a DC component*, it can be shown that,

$$y_0 = 0.$$

$$A_n = 0.$$

$$B_n = \frac{2Y}{\pi n} [1 - (-1)^n]$$

$$\text{For even } n, (\text{Even harmonics}), B_n = 0. \quad (4)$$

$$\text{For odd } n, (\text{Odd Harmonics}), B_n = \frac{4Y}{n\pi} \quad (5)$$

5.1 ADALINE

Artificial Neural Networks have proven their efficacy in many applications where conventional algorithms require more computational time and complexity. One class of model called ADALINE (**Adaptive Linear Element**) uses the learning rule proposed by Widrow and Hoff [21]. ADALINE has the following advantages: better convergence than perceptron, does not require any memory and on-line. Using this model, a nonsinusoidal wave form may be expressed as a linear combination of the fundamental and harmonics with different amplitudes.

The general form of a non sinusoidal waveform is,

$$y(t) = \sum_{n=1}^N A_n \sin(n\omega t + \phi_n) \quad (6)$$

where the A_n and ϕ_n are the amplitude and phase of the harmonics, respectively, N is the maximum order of harmonics up to which the input waveform has to be



resolved and t is the time instant of measurement. The discrete version of equation (7) is

$$y(k) = \sum_{n=1}^N A_n \sin\left(\frac{2\pi nk}{N_s} + \phi_n\right) \quad (7)$$

$$= \sum_{n=1}^N A_n \sin\left(\frac{2\pi nk}{N_s}\right) \cos\phi_n + \sum_{n=1}^N A_n \cos\left(\frac{2\pi nk}{N_s}\right) \sin\phi_n \quad (8)$$

where N_s is the sample rate given by $N_s = f_s / f_1$ and k is the instant of iteration.

The ADALINE model for measuring harmonics present in the sampled input waveform $y(k)$ is given in figure 1.

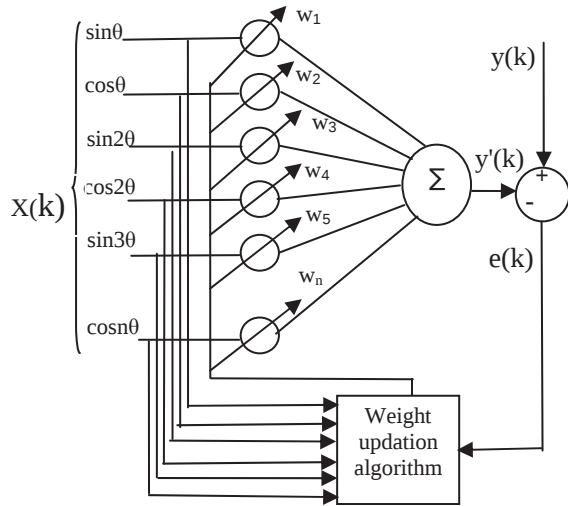


Figure 1. The ADALINE model

The training vector $X(k)$ is formed as,

$$X(k) = [\sin(2\pi k/N_s) \cos(2\pi k/N_s) \sin(4\pi k/N_s) \cos(4\pi k/N_s) \dots \sin(2N\pi k/N_s) \cos(2N\pi k/N_s)]^T \quad (9)$$

In figure 1, $X(k)$ is shown as $[\sin\theta \cos\theta \sin2\theta \cos2\theta \dots \sin n\theta \cos n\theta]$ where $\theta = 2\pi k/N_s$. The initial weight vector $W = [w_1 w_2 w_3 \dots w_n]^T$ is randomly chosen to be small values.

The weight vector $W(k)$ is updated using Widrow-Hoff rule as,

$$W(k+1) = W(k) + \frac{\alpha e(k) X(k)}{X^T(k) X(k)} \quad (10)$$

Where $W(k) = [W_1(k) W_2(k) W_3(k) \dots W_{2N}(k)]^T$, α is the learning parameter, $X(k)$ is the input training vector at time k , $e(k) = y(k) - y'(k)$ is the error at time k , $y(k)$ is the actual amplitude of input signal at time k and $y'(k)$ is the estimated amplitude at time k .

On the k^{th} iteration, the error is, $e(k) = y(k) - y'(k)$

When the ADALINE network has perfectly learnt the input signal (ie. convergence is reached), this error will be zero. Then,

$$y(k) = W_0^T X(k) \quad (11)$$

where W_0^T is the weight vector after convergence. Thus the model can predict the input nonsinusoidal waveform correctly. The convergence of the algorithm has been proved by considering the Lyapunov's energy function [14].

After the tracking error converges to zero, the Fourier coefficients corresponding to various harmonics are obtained from the weight vector by,

$$A_N = [A_1 \sin\phi_1, A_1 \cos\phi_1, \dots, A_N \sin\phi_N, A_N \cos\phi_N] \quad (12)$$

The amplitude and phase of the n^{th} harmonic is given by,

$$A_n = \sqrt{W_0^2(2n-1) + W_0^2(2n)} \quad (13)$$

$$\text{and } \phi_n = \tan^{-1} \left[\frac{W_0(2n-1)}{W_0(2n)} \right] \quad (14)$$

6. Real Time Computation of THD Using FPGA

In [22] the ADALINE algorithm for computation of THD was implemented on a DSP.

In this paper, the ADALINE algorithm has been implemented on a Virtex-II FPGA of Xilinx. The Virtex-II reference board utilizes the Xilinx 40K/1M gate Virtex-II (XC2V40 or XC2V1000) device. The FPGA does not contain an in-built Analog to Digital Converter (ADC). Hence an external ADC was used to convert the current signal. The ADC used is LTC1407-1, 12-bit, 3Msps with two 1.5Msps simultaneously sampled differential inputs.

The block diagram of the experimental setup is shown in figure 2.

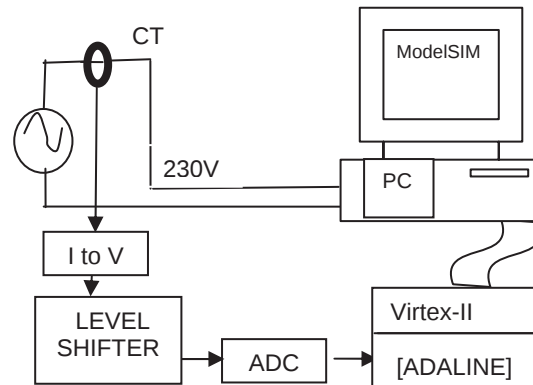


Figure 2 Block diagram of the experimental setup.

The VHDL code generated in ModelSim used real values, but implementation in FPGA does not permit real values. Hence all values were converted into integers and then implemented in Virtex-II FPGA kit. Here deviation increased to a certain extent.

A training vector of 80 x 100 samples was formed. The number of rows in the training vector was chosen as 80 so as to compute up to 40th order harmonics. The number of columns in the training vector was chosen as 100 to accommodate 100 samples taken over one cycle of 50Hz. The training vector was computed *a priori* and stored in the memory.

The load current waveform of a Personal Computer, which is a non-linear load, was acquired by a signal processing circuit. This circuit comprises of a current



transformer connected in series with the load, a current-to-voltage converter (I-V) and an amplifier. The output of the amplifier is fed to the ADC. The digital data from the ADC is read by the FPGA. A zero crossing detector circuit was used to initiate the acquisition of samples. 100 samples at an interval of $200\mu\text{s}$, spanning over one complete cycle of 20ms, were obtained. The implementation code is written in VHDL.

From the samples acquired by FPGA the load current waveform and supply voltage were reconstructed using a spreadsheet program. These are shown in figure 3. The figure shows that the voltage is sinusoidal whereas the current wave form is not sinusoidal.

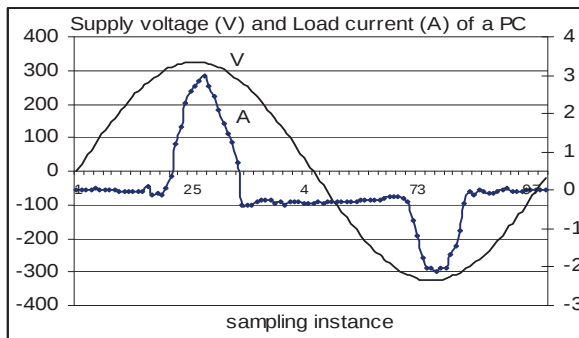


Figure 3. Supply voltage and load current waveforms of a PC.

7. Results And Discussion

The 100 sampled values of the load current signal $y(k)$ were applied to the ADALINE network. It was observed that the ADALINE network converges in 10 epochs. The convergence of error for the first ten harmonics, $A_1, A_2, \dots, A_{10}$ (out of $A_1 - A_{25}$) towards the final value is shown in figure 4. The convergence of weight values ($W_1, W_2, \dots, W_{50}$) of ADALINE network towards their final values is shown in figure 5. For clarity, the graph shows only the first 10 values viz. $W_1, W_2, \dots, W_{10}$.

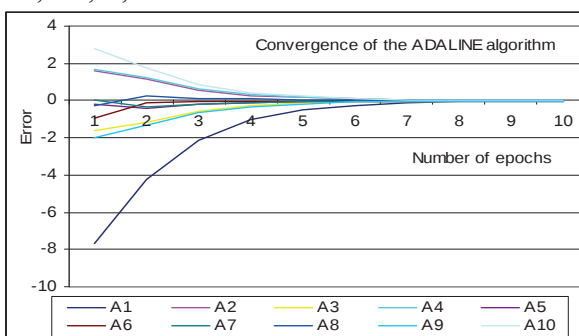


Figure 4. Plot of the epochs versus error .

The harmonics contents in the load current waveform were computed using (13) and are given in Table 2. The table also lists the values obtained from a standard measuring equipment, namely Power Quality Analyser (PQA) model CA 8332 of Chauvin-Arnoux. Since the final objective of this work is to compute the THD, to simplify the calculations, only squares of the amplitudes of A_n were calculated using VHDL.

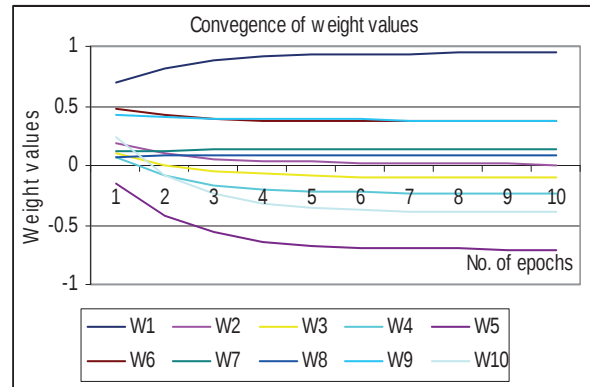


Figure 5. Convergence of weight values within 10 epochs.

Table 2. Comparison of results from the VHDL and PQA

Harmonics order A_n	Values of A_n^2 computed from readings of PQA	Values of A_n^2 Computed by VHDL	Error %
1	4678.56	4618.7	1.30
2	327.61	342.17	4.26
3	3340.84	3267.96	2.23
4	134.56	132.58	1.49
5	1576.09	1543.81	2.09
6	22.09	25.36	12.89
7	380.25	388.11	2.03
8	6.76	7.45	9.26
9	29.16	18.38	58.65
10	21.16	19.35	9.35
11	17.64	26.61	33.71
12	16.81	17.29	2.78
13	18.49	32.32	42.79
14	7.29	8.10	10.0
15	2.25	2.39	5.86
16	16.0	26.13	38.77
17	4.0	6.72	40.47
18	9.61	18.03	46.70
19	5.29	9.82	46.13
20	2.25	5.02	55.18
21	2.25	3.76	40.16
22	2.25	7.65	70.59
23	0.25	1.42	82.39
24	2.25	8.95	74.86
THD	127.07%	128.16%	0.85%

The comparison shows that the ADALINE algorithm implemented on a FPGA using VHDL produces comparable results to that of an FFT algorithm.

The advantage of the proposed implementation is that 100 sample only are needed from a 20ms period compared to 1000 samples from a period of 200ms.

The value of total harmonic distortion was calculated using (15) separately with the data obtained from PQA and VHDL.

Value of THD using PQA = 127.0%



Value of THD using VHDL =128.16%

The number of multiplications and additions required for implementing the FFT algorithm and Adaline algorithm are listed in table 3.

Table 3. Comparison of Computational Complexity between the FFT and proposed ADALINE algorithms

Number of Operation	For 1024 point FFT algorithm	For ADALINE algorithm requiring 10 epochs to converge
Multiplications	102,400	76,048
Additions	51,200	50,048

8. Conclusion

The paper presents an implementation methodology to adopt the ADALINE algorithm on a FPGA to calculate the THD in real time load current waveforms. When compared with the conventional FFT algorithm, the ADALINE algorithm requires only 100 data samples, at intervals of 200 μ s, spanning over one cycle only. The algorithm also converges in 3 epochs, making it faster than the FFT algorithm. The accuracy of the results is also comparable to that of standard Power Quality Analyser. As can be seen in the Table 3, the proposed ADALINE algorithm requires less number of operations than the conventional of FFT algorithm. It is shown that the proposed methodology is efficient for stand alone hardware implementation. The measurement of harmonics in real time also makes this method suitable for adopting in filter circuits. Faster performance with lower power consumption can be achieved by implementing this algorithm using any technique like fine grain pipelining, distributed arithmetic algorithm, variable precision fine grain architecture, dynamic data range detection, partial guarded computation, spurious power suppression technique, etc.

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Decentralized Observer Based H_∞ Decentralized Control for Interconnected Systems

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Abstract

This paper presents an approach of H_∞ decentralized tracking control design with decentralized observer for large scale interconnected systems. The proposed control structure is developed to guarantee the asymptotic stability of the global system, to improve its performances by using a modified H_∞ criterion and to ensure the tracking of a reference model with the minimization of the external disturbances effect. Moreover, the synthesized controller includes a decentralized observer which allows the reconstruction of the non measurable state variables of each subsystem. The determination of the control and observer gains is derived by means of an optimization problem which is formulated in terms of linear matrix inequalities (LMIs). Simulation results on a three interconnected machine power system are presented to highlight the performances of the proposed control scheme. **Keywords:** *Large scale systems, Interconnected power systems, Decentralized observer, H_∞ decentralized tracking control, LMI approach.*

1 Introduction

Recent years have seen significant progress in the field of control of both linear and nonlinear large scale interconnected systems. The study of this kind of systems is motivated by a number of practical applications ranging from power networks, aerospace, transportation, economics, management and others.

Large scale interconnected systems are often modeled by dynamic equations composed of the interconnection of lower-dimensional subsystems. These systems are widely distributed in space, thus information transfer among subsystems may be very difficult. Therefore, decentralized control schemes are considered as effective methods to deal with large scale interconnected systems since they avoid the transfer information between subsystems [2, 25, 20, 7, 21].

Decentralized control for large scale interconnected systems has received considerable interest in the literature because it presents an efficient means for designing control algorithms. Numerous works of research have been developed and reported concerning this topic such as the state feedback decentralized control, the robust decentralized control and the decentralized output feedback control [19, 22, 9, 11, 10, 12, 24].

The ability to control an interconnected system by decentralized state feedback depends crucially upon the availability of states at each subsystem. In most practical cases, the complete state measurements are not available at each individual subsystem. Consequently, the state observation techniques may be required to reconstruct the non measurable subsystem states [15, 5, 3].

Several works were focused on the observer design for large scale interconnected systems. The interconnected observer design was proposed in [8, 13, 14]. In this structure, the interconnection terms were used in the observer scheme. In addition, decentralized observer and decentralized observer based state feedback control, where there is no information transfer between local observers, was also given [17, 18, 16].

However, the problem of decentralized observer based feedback control for interconnected systems is not obvious since the separation principle may not be applicable in this situation. Thus, it is necessary to consider simultaneously the problem of control and observer design to ensure the global asymptotic stabilization [1, 16].

The problem of decentralized estimated state feedback control has been studied in this work not only to obtain the asymptotic stability but also to guarantee a high performance control law with some prescribed constraints for the interconnected system. Thus, the main contribution of this paper is to design a decentralized observer based H_∞ decentralized model reference tracking control to ensure the asymptotic sta-



bilization and H_∞ performances of the global system allowing the minimization of the external disturbances effect. The determination of the control and observer gains is derived by means of an optimization problem which is formulated in terms of linear matrix inequalities (LMIs).

An outline of this paper is as follows: the problem formulation is introduced in section (2). Then in section (3), the stability study of the augmented system is presented, and the decentralized observer based H_∞ decentralized tracking control approach is designed in terms of LMI optimization problem. While, section (4) is devoted in the first part to the multi-machine power system modeling and in the second part to numerical simulations, conducted on three interconnected machine power system, to illustrate the validity and the effectiveness of the proposed approach. Finally, some conclusions are provided in section (5).

2 Problem formulation

The objective is to design a totally decentralized observer based H_∞ decentralized tracking control approach for large scale interconnected systems that stabilizes the states of the overall system.

The class of large scale interconnected systems considered in this work can be characterized by the interconnection of N linear subsystems as follows:

$$\begin{cases} \dot{x}_i(t) = A_i x_i(t) + B_i u_i(t) + \sum_{j=1, j \neq i}^N H_{ij} x_j(t) + w_i(t) \\ y_i(t) = C_i x_i(t) + v_i(t) \end{cases} ; \quad i = 1, \dots, N \quad (1)$$

where:

- $x_i(t) \in R^{n_i}$ is the state vector of the i^{th} subsystem,
- $u_i(t) \in R^{m_i}$ is the control vector of the i^{th} subsystem,
- $y_i(t) \in R^{p_i}$ is the output vector of the i^{th} subsystem,
- $w_i(t)$ is the external disturbance of the i^{th} subsystem,
- $v_i(t)$ is the measurement noise of the i^{th} subsystem,
- $A_i \in R^{n_i \times n_i}$, $B_i \in R^{n_i \times m_i}$, $C_i \in R^{p_i \times n_i}$, are respectively the state matrix, the input matrix and the output matrix of each subsystem,
- (A_i, B_i) is controllable,
- (A_i, C_i) is observable,
- $H_{ij} \in R^{n_i \times n_j}$ are constant matrix representing the linear interconnection between the two subsystems i and j .

The global system can then be modeled by the following state representation:

$$\begin{cases} \dot{x}(t) = Ax(t) + Bu(t) + Hx(t) + w(t) \\ y(t) = Cx(t) + v(t) \end{cases} \quad (2)$$

where:

- $x^T(t) = x^T = [x_1^T, x_2^T, \dots, x_N^T]$ is the state vector of

the global system,

- $u^T(t) = u^T = [u_1^T, u_2^T, \dots, u_N^T]$ is the control vector of the global system,
- $y^T(t) = y^T = [y_1^T, y_2^T, \dots, y_N^T]$ is the output vector of the global system,
- $w^T(t) = w^T = [w_1^T, w_2^T, \dots, w_N^T]$ is the disturbance vector of the global system,
- $v^T(t) = v^T = [v_1^T, v_2^T, \dots, v_N^T]$ is the noise vector of the global system,
- $A = \text{diag}(A_i)$, $B = \text{diag}(B_i)$ and H is the interconnection matrix satisfying $H = [H_{ij}]$ with $H_{ii} = 0$.
- The pair (A, B) is controllable and the pair (A, C) is observable. This property is the direct result of each subsystem being controllable and observable.

Consider a reference model for the i^{th} subsystem as follows:

$$\dot{x}_{ri}(t) = A_{ri} x_{ri}(t) + r_i(t) \quad (3)$$

where:

- $x_{ri}(t) \in R^{n_i}$ denotes the reference state vector of the i^{th} subsystem. It is assumed that $x_{ri}(t)$ represents a desired trajectory for $x_i(t)$ for all $t \geq 0$.
- $A_{ri} \in R^{n_i \times n_i}$ denotes a specific asymptotically stable matrix,
- $r_i(t) \in R^{n_i}$ denotes a bounded reference input.

The reference model for the global system is designed by the following equation:

$$\dot{x}_r(t) = A_r x_r(t) + r(t) \quad (4)$$

where:

- $x_r^T(t) = x_r^T = [x_{r1}^T, x_{r2}^T, \dots, x_{rN}^T]$,
- $r^T(t) = r^T = [r_1^T, r_2^T, \dots, r_N^T]$ denotes the bounded reference input of the global system,
- $A_r = \text{diag}(A_{ri})$.

The local control law of each subsystem is given by:

$$u_i = K_i [x_{ri}(t) - \hat{x}_i(t)] \quad (5)$$

where:

- K_i is the control gain of the i^{th} subsystem,
- $\hat{x}_i = \hat{x}_i(t)$ is the observed state vector described by the following equations:

$$\begin{cases} \dot{\hat{x}}_i(t) = A_i \hat{x}_i(t) + B_i u_i(t) + L_i [y_i(t) - \hat{y}_i(t)] \\ \hat{y}_i(t) = C_i \hat{x}_i(t) \end{cases} ; \quad i = 1, \dots, N \quad (6)$$

with L_i the observation gain of the i^{th} subsystem.

The error observation, $e_i(t) = e_i$, between the real state and the observed one:

$$e_i = x_i - \hat{x}_i \quad (7)$$

is described by the following equation:

$$\dot{e}_i = [A_i - L_i C_i] e_i - L_i v_i + \sum_{j=1, j \neq i}^N H_{ij} x_j + w_i \quad (8)$$



The observer of the global system can be expressed by the following form:

$$\begin{cases} \dot{\hat{x}} = A\hat{x} + Bu + L[y - \hat{y}] \\ \hat{y} = C\hat{x} \end{cases} \quad (9)$$

with $\hat{x}^T = [\hat{x}_1^T, \hat{x}_2^T, \dots, \hat{x}_N^T]$ and $L = \text{diag}(L_i)$ is the observation gain matrix.

Notice that the state estimation scheme of the global interconnected system, composed by N local observers, is completely decentralized since each local observer uses only the input and the output of the considered subsystem. This decentralized observer had not been considered in previous works related to the subject of decentralized control as the approach proposed in [6] which had considered the H_∞ decentralized fuzzy model reference tracking control design using interconnected observers since the interconnection terms ($H_{ij}x_j(t)$) were conserved in the used observer structure. Thus, the contribution of our work is to avoid the interconnection of the observer and to synthesize a control structure based on a completely decentralized observer.

In order to design the tracking control law of the global system (2), one considers the following block diagonal control:

$$u = K[x_r - \hat{x}] \quad (10)$$

where $K = \text{diag}(K_i)$ is the control gain matrix.

The global observation error system, $e = x - \hat{x}$, can be developed, by using (2) and (9), in the following form:

$$\dot{e} = [A - LC]e - Lv + Hx + w \quad (11)$$

Substituting the control law (10) in the global system (2), the N interconnected subsystems can be modeled by the following state representation:

$$\begin{cases} \dot{x} = (A + H - BK)x + BKx_r + BKe + w \\ y = Cx + v \end{cases} \quad (12)$$

Therefore, the augmented system including the overall system (12), the global reference model (4) and the global observation error system (11), is presented in the following form:

$$\dot{\tilde{x}} = \tilde{A}\tilde{x} + \tilde{B}\tilde{w} \quad (13)$$

where:

$$\tilde{x} = \begin{bmatrix} x \\ x_r \\ e \end{bmatrix}, \quad \tilde{A} = \begin{bmatrix} A + H - BK & BK & BK \\ 0 & A_r & 0 \\ H & 0 & A - LC \end{bmatrix},$$

$$\tilde{B} = \begin{bmatrix} I & 0 & 0 \\ 0 & I & 0 \\ I & 0 & -L \end{bmatrix}, \quad \tilde{w} = \begin{bmatrix} w \\ r \\ e \end{bmatrix}.$$

The problem is to find a way to obtain the control gain matrix $K = \text{diag}(K_i)$ and the observation

gain matrix $L = \text{diag}(L_i)$ which can achieve the stability and the H_∞ tracking control performances of the augmented system (13) in same time.

3 Stability and H_∞ tracking control performances of the augmented system

This section is devoted to analyze the stability and to improve the H_∞ tracking control performances of the augmented system (13).

In order to study the stability, which is the most important issue in the control system, we consider a quadratic positive definite Lyapunov function as follows:

$$V(x, x_r, e) = x^T P_c x + x_r^T P_r x_r + e^T P_o e \quad (14)$$

which can be also written as:

$$V(\tilde{x}) = \tilde{x}^T P \tilde{x} \quad (15)$$

with $P = \text{diag}(P_c, P_r, P_o)$ and $P_c = \text{diag}(P_{ci})$, $P_r = \text{diag}(P_{ri})$, $P_o = \text{diag}(P_{oi})$ are positive definite symmetric matrices.

The time derivative of $V(x, x_r, e)$ expressed by:

$$\dot{V}(\tilde{x}) = \dot{\tilde{x}}^T P \tilde{x} + \tilde{x}^T P \dot{\tilde{x}} \quad (16)$$

must be negative definite if we consider the autonomous system:

$$\dot{\tilde{x}} = \tilde{A}\tilde{x} \quad (17)$$

Thus, a sufficient condition for the asymptotic stability of (17) is given by the following inequality:

$$\tilde{x}^T [\tilde{A}^T P + P \tilde{A}] \tilde{x} \leq 0 \quad (18)$$

which leads to:

$$\begin{bmatrix} X & P_c B K & P_c B K + H^T P_o \\ (B K)^T P_c & A_r^T P_r + P_r A_r & 0 \\ (B K)^T P_c + P_o H & 0 & Y \end{bmatrix} \leq 0 \quad (19)$$

with $X = (A + H - BK)^T P_c + P_c (A + H - BK)$ and $Y = (A - LC)^T P_o + P_o (A - LC)$.

In the purpose to design the H_∞ decentralized tracking control based on the decentralized observer (9) for the augmented system (13), let us consider the H_∞ tracking performance related to the tracking error $e_{ri} = x_{ri}(t) - x_i(t)$ and the observation error $e_i = x_i(t) - \hat{x}_i(t)$ by the following constraint:

$$\frac{\sum_{i=1}^N \int_0^{tf} e_{ri}^T Q_i e_{ri} dt + \sum_{i=1}^N \int_0^{tf} e_i^T Q_{ei} e_i dt}{\int_0^{tf} \tilde{w}^T \tilde{w} dt} \leq \rho^2 \quad (20)$$



which can be written as follows:

$$\sum_{i=1}^N \int_0^{t_f} e_{ri}^T Q_i e_{ri} dt + \sum_{i=1}^N \int_0^{t_f} e_i^T Q_{ei} e_i dt = \int_0^{t_f} \tilde{x}^T \tilde{Q} \tilde{x} dt \leq \rho^2 \int_0^{t_f} \tilde{w}^T \tilde{w} dt \quad (21)$$

where:

- t_f is the terminal time of control,
- ρ is the prescribed attenuation level,
- Q and Q_e are symmetric positive definite matrices described by:

$$Q = \text{diag}(Q_i); \quad Q_e = \text{diag}(Q_{ei}); \quad i = 1, \dots, N$$

with Q_i et Q_{ei} are symmetric positive definite weighting matrices.

- $\tilde{Q}$, designed from (21), has the form:

$$\tilde{Q} = \begin{bmatrix} Q & -Q & 0 \\ -Q & Q & 0 \\ 0 & 0 & Q_e \end{bmatrix}$$

If the initial condition is also considered, the inequality (21) can be modified as:

$$\int_0^{t_f} \tilde{x}^T \tilde{Q} \tilde{x} dt \leq \tilde{x}^T(0) P \tilde{x}(0) + \rho^2 \int_0^{t_f} \tilde{w}^T \tilde{w} dt \quad (22)$$

If the attenuation level ρ is minimized, then the H_∞ tracking performances (21) are reduced as small as possible.

The development of the inequality (22) according to the state representation of the augmented system (13) gives:

$$\begin{aligned} \int_0^{t_f} \tilde{x}^T \tilde{Q} \tilde{x} dt &= \tilde{x}^T(0) P \tilde{x}(0) - \tilde{x}^T(t_f) P \tilde{x}(t_f) \\ &+ \int_0^{t_f} \left\{ \tilde{x}^T \tilde{Q} \tilde{x} + \frac{d[\tilde{x}^T P \tilde{x}(t)]}{dt} \right\} dt \leq \tilde{x}^T(0) P \tilde{x}(0) \\ &+ \int_0^{t_f} \left\{ \tilde{x}^T \tilde{Q} \tilde{x} + \dot{\tilde{x}}^T P \tilde{x} + \tilde{x}^T P \dot{\tilde{x}} \right\} dt \end{aligned} \quad (23)$$

then the H_∞ tracking performance (23) becomes:

$$\int_0^{t_f} \tilde{x}^T \tilde{Q} \tilde{x} dt \leq \tilde{x}^T(0) P \tilde{x}(0) \quad (24)$$

$$\begin{aligned} &+ \int_0^{t_f} \left\{ \tilde{x}^T \tilde{Q} \tilde{x} + \dot{\tilde{x}}^T P \tilde{x} + \tilde{x}^T P \dot{\tilde{x}} + \rho^2 \tilde{w}^T \tilde{w} - \rho^2 \tilde{w}^T \tilde{w} \right\} dt \\ &\leq \tilde{x}^T(0) P \tilde{x}(0) + \int_0^{t_f} \left\{ \tilde{x}^T A_c \tilde{x} + B_c^T \tilde{w} + \rho^2 \tilde{w}^T \tilde{w} - \rho^2 \tilde{w}^T \tilde{w} \right\} dt \end{aligned}$$

with $A_c = \tilde{A}^T P + P \tilde{A} + \tilde{Q}$ and $B_c = \tilde{w}^T \tilde{B}^T P \tilde{x} + \tilde{x}^T P \tilde{B} \tilde{w}$.

Finally, it yields the following inequality:

$$\int_0^{t_f} \tilde{x}^T \tilde{Q} \tilde{x} dt \leq \tilde{x}^T(0) P \tilde{x}(0) \quad (25)$$

$$+ \int_0^{t_f} \left\{ \begin{bmatrix} \tilde{x} \\ \tilde{w} \end{bmatrix}^T \begin{bmatrix} A_c & P \tilde{B} \\ \tilde{B}^T P & -\rho^2 I \end{bmatrix} \begin{bmatrix} \tilde{x} \\ \tilde{w} \end{bmatrix} + \rho^2 \tilde{w}^T \tilde{w} \right\} dt$$

For the linear augmented system (13), if $P = P^T > 0$ is the solution of the following matrix inequality:

$$\begin{bmatrix} \tilde{A}^T P + P \tilde{A} + \tilde{Q} & P \tilde{B} \\ \tilde{B}^T P & -\rho^2 I \end{bmatrix} \leq 0 \quad (26)$$

then the augmented interconnected system is stable in the Lyapunov meaning and the H_∞ tracking control based on decentralized observer is guaranteed for a prescribed ρ .

The development of the previous inequality (26) leads to:

$$\begin{bmatrix} X_{11} & X_2 & X_3 & P_c & 0 & 0 \\ X_2^T & X_{22} & 0 & 0 & P_r & 0 \\ X_3^T & 0 & X_{33} & P_o & 0 & -P_o L \\ P_c & 0 & P_o & -\rho^2 I & 0 & 0 \\ 0 & P_r & 0 & 0 & -\rho^2 I & 0 \\ 0 & 0 & -L^T P_o & 0 & 0 & -\rho^2 I \end{bmatrix} \leq 0 \quad (27)$$

with:

- $X_{11} = (A + H - BK)^T P_c + P_c (A + H - BK) + Q$,
- $X_{22} = A_r^T P_r + P_r A_r + Q$,
- $X_{33} = (A - LC)^T P_o + P_o (A - LC) + Q_e$,
- $X_2 = P_c BK - Q$,
- $X_3 = P_c BK + H^T P_o$.

To obtain better tracking performances, the attenuation level ρ must be reduced as small as possible. Indeed, the H_∞ decentralized observer based tracking control problem can be formulated by the following minimization problem:

$$\begin{cases} \text{minimize } \rho^2 \\ \text{subject to} \\ P_c = P_c^T > 0, \quad P_r = P_r^T > 0, \quad P_o = P_o^T > 0 \\ \text{and inequality} \end{cases} \quad (28)$$

(27)

In the minimization problem described by (28), variables are $P_c, P_r, P_o, Q, Q_e, K, L$ and ρ . Since the inequality (27) contains coupled terms of matrix variables (P_c, K) and (P_o, L) , the minimization problem (28) results in a bilinear matrix inequality.



One must then find a means to transform the inequality (27) into a form which is affine in the unknown variables. To achieve this, one introduce the following variables:

$$M_{co} = P_c B K \quad , \quad M_{ob} = P_o L. \quad (29)$$

The matrices M_{co} and M_{ob} were simultaneously extracted from the resolution of the minimization problem (28). The observation gain block diagonal matrix $L = \text{diag}(L_i)$ can be calculated from M_{ob} as:

$$L = P_o^{-1} M_{ob}. \quad (30)$$

While, The control gain matrix $K = \text{diag}(K_i)$ can be obtained from M_{co} only in the case when the control matrix B is invertible, that is:

$$K = B^{-1} P_c^{-1} M_{co}. \quad (31)$$

However, the invertibility of the diagonal matrix $B = \text{diag}(B_i)$ requires that the matrix B_i be invertible, which is too restrictive. The following treats the solution to the case when B_i is not invertible. To transform the inequality (27) to a form which is affine in the unknown variables, we can multiply (27) on the left and on the right hand by the matrix:

$$M = \begin{bmatrix} Y_c & 0 & 0 & 0 & 0 & 0 \\ 0 & I & 0 & 0 & 0 & 0 \\ 0 & 0 & I & 0 & 0 & 0 \\ 0 & 0 & 0 & I & 0 & 0 \\ 0 & 0 & 0 & 0 & I & 0 \\ 0 & 0 & 0 & 0 & 0 & I \end{bmatrix} \quad (32)$$

with $Y_c = P_c^{-1}$. Then, the inequality (27) becomes:

$$\begin{bmatrix} M_{11} & M_2 & M_3 & I & 0 & 0 \\ M_2^T & M_{22} & 0 & 0 & P_r & 0 \\ M_3^T & 0 & M_{33} & P_o & 0 & -F \\ I & 0 & P_o & -\rho^2 I & 0 & 0 \\ 0 & P_r & 0 & 0 & -\rho^2 I & 0 \\ 0 & 0 & -F^T & 0 & 0 & -\rho^2 I \end{bmatrix} \leq 0 \quad (33)$$

with:

- $M_{11} = A_g Y_c + Y_c A_g^T - (BR)^T - BR + Y_c Q Y_c$,
- $M_{22} = X_{22}$,
- $M_{33} = A^T P_o + P_o A - (FC)^T - FC + Q_e$,
- $M_2 = BK - Y_c Q$,
- $M_3 = BK + Y_c H^T P_o$,
- $A_g = A + H$
- $R = K Y_c$,
- $F = P_o L$.

Therefore, the H_∞ decentralized observer based tracking control problem can be reformulated as the following optimization problem:

$$\begin{cases} \text{minimize } \rho^2 \\ \text{subject to} \\ Y_c = Y_c^T > 0, \quad P_r = P_r^T > 0, \quad P_o = P_o^T > 0 \\ \text{and inequality} \end{cases} \quad (34)$$

The control and the observer problems can be decoupled and can be solved by two Step procedures which solve the control parameters first and then solve the observer parameters.

We first compute the matrices Y_c and R from the following matrix inequality:

$$\begin{bmatrix} A_g Y_c + Y_c A_g^T - (BR)^T - BR & Y_c \\ Y_c & -Q^{-1} \end{bmatrix} \leq 0 \quad (35)$$

The control gain matrix is obtained from the step1 as:

$$K = R Y_c^{-1} \quad (36)$$

Using the parameters obtained from The first step, the second step is considered to find P_o , P_r and F by solving the following minimization problem:

$$\begin{cases} \text{minimize } \rho^2 \\ \text{subject to } P_r = P_r^T > 0, \quad P_o = P_o^T > 0 \\ \text{and inequality} \end{cases} \quad (37)$$

Then the observer gain matrix is obtained as follows:

$$L = P_o^{-1} F \quad (38)$$

4 Application of the proposed control approach to multi-machine power system

4.1 Multi-machine power system non-linear model

An N machine power system with steam valve control can be described by the interconnection of N subsystems. The nonlinear mathematical model of each subsystem is described by the following equations [23, 4]:

$$\begin{cases} \dot{x}(t) = A_i x_i(t) + B_i u_i(t) + h_i(t, x(t)) \\ y_i(t) = C_i x_i(t) \quad ; \quad i = 1, \dots, N \end{cases} \quad (39)$$

with:

$$h_i(t, x(t)) = \sum_{j=1, j \neq i}^N p_{ij} G_{ij} g_{ij}(x_i(t), x_j(t))$$

is a nonlinear function characterizing the interconnection between i^{th} and j^{th} machine and $x_i(t)$ is the state vector defined by:

$$x_i(t)^T = [\Delta \delta_i(t) \quad \omega_i(t) \quad \Delta P_{mi}(t) \quad \Delta X_{ei}(t)]$$



where:

- $\Delta\delta_i(t) = \delta_i(t) - \delta_{i0}$,
- $\Delta P_{m_i}(t) = P_{m_i}(t) - P_{m_{i0}}$,
- $\Delta X_{e_i}(t) = X_{e_i}(t) - X_{e_{i0}}$,
- $u_i(t)$ is the control vector, $u_i(t) = P_{ci} - P_{m_{i0}}$,
- $y_i(t)$ is the output vector, $y_i(t) = \Delta\delta_i(t)$,

$$A_i = \begin{bmatrix} 0 & 1 & 0 & 0 \\ 0 & -\frac{D_i}{2H_i} & \frac{\omega_0}{2H_i} & 0 \\ 0 & 0 & -\frac{1}{T_{m_i}} & \frac{K_{m_i}}{T_{m_i}} \\ 0 & -\frac{K_{e_i}}{T_{e_i} R_i \omega_0} & 0 & -\frac{1}{T_{e_i}} \end{bmatrix}, \quad B_i = \begin{bmatrix} 0 \\ 0 \\ 0 \\ \frac{1}{T_{e_i}} \end{bmatrix},$$

$$C_i = [1 \ 0 \ 0 \ 0],$$

$$G_{ij}^T = \begin{bmatrix} 0 & -\frac{\omega_0 E'_{qi} E'_{qj} B_{ij}}{2H_i} & 0 & 0 \end{bmatrix},$$

$$g_{ij}(x_i(t), x_j(t)) = \sin(\delta_i(t) - \delta_j(t)) - \sin(\delta_{i0} - \delta_{j0}).$$

with:

- $\delta_i(t)$ Rotor angle for i^{th} machine;
- $\omega_i(t)$ Relative speed for i^{th} machine;
- $P_{m_i}(t)$ Mechanical power for i^{th} machine;
- P_{ci} Power control input of the i^{th} machine;
- $X_{e_i}(t)$ Steam valve opening for i^{th} machine;
- H_i Inertia constant for the i^{th} machine, in second;
- D_i Damping coefficient for the i^{th} machine, in pu ;
- T_{m_i} Time constant for i^{th} machine's turbine, in second;
- K_{m_i} Gain of i^{th} machine's turbine;
- T_{e_i} Time constant of the i^{th} machine's speed governor, in second;
- K_{e_i} Gain of the i^{th} machine's speed governor;
- R_i Regulation constant of the i^{th} machine, in pu ;
- B_{ij} Nodal susceptance between i^{th} and j^{th} machines, in pu ;
- ω_0 Synchronous machine speed, en radian/s;
- E'_{qi} Internal transient voltage for i^{th} machine, in pu , assumed to be constant;
- E'_{qj} Internal transient voltage for j^{th} machine, in pu , assumed to be constant;
- p_{ij} Constant of either 1 or 0 ($p_{ij} = 0$ means that i^{th} machine has no connection with j^{th} machine);

δ_{i0} , $P_{m_{i0}}$ and $X_{e_{i0}}$ are the nominal values of $\delta_i(t)$, $P_{m_i}(t)$ and $X_{e_i}(t)$.

4.2 Linearized multi-machine power system model

For little variations of the load angle δ_i of the i^{th} machine around an equilibrium state δ_{i0} , the term $h_i(t, x(t))$ in (39) can be presented by the following linear form:

$$h_i(t, x(t)) = \sum_{j=1, j \neq i}^N p_{ij} G_{ij} (\Delta\delta_i(t) - \Delta\delta_j(t)) \quad (40)$$

Then, the linearized model of each subsystem can be presented by the following system:

$$\begin{cases} \dot{x}_i(t) = A_i x_i(t) + B_i u_i(t) + \sum_{j=1, j \neq i}^N H_{ij} x_j(t) \\ y_i(t) = C_i x_i(t) \end{cases} ; \quad i = 1, \dots, N \quad (41)$$

The N interconnected machine power system can be modeled, as shown in the equation (2), by the following state representation:

$$\begin{cases} \dot{x} = Ax + Bu + Hx + w \\ y = Cx + v \end{cases} \quad (42)$$

4.3 Numerical simulations

To study the availability and the effectiveness of the proposed control approach, we consider three interconnected machine power system characterized by the parameters summarized in the following table [23]:

	Machine 1	Machine 2	Machine 3
$T'_{d0}(pu)$	6.9	7.96	7.96
$H(s)$	4	5.1	5.1
$D(pu)$	5	3	3
$T_m(s)$	0.35	0.35	0.35
$T_e(s)$	0.1	0.1	0.1
R	0.05	0.05	0.05
K_m	1	1	1
K_e	1	1	1
$\omega_0(rad/s)$	314.159	314.159	314.159

Table 1: Parameters of the three interconnected machine power system

The state matrices of the stable reference models are given by:

$$A_{ri} = \begin{bmatrix} 0 & 1 & 0 & 0 \\ -100 & -101 & 0 & 0 \\ 0 & 0 & -50 & -2 \\ 0 & -2 & 0 & -10 \end{bmatrix}, \quad i = 1, 2, 3$$

Using the above parameters, the decentralized control gain matrices and the observation gain matrices were extracted from the two following stages.

In the first step, the resolution of the LMI problem (??) permits the calculus of the decentralized control gain $K = \text{diag}(K_i)$, $i = 1, 2, 3$, as follows:

$$\begin{aligned} K_1 &= [5.6341 \quad 3.9697 \quad 20.4060 \quad 1.0750] \\ K_2 &= [12.9935 \quad 6.1510 \quad 22.1360 \quad 1.4951] \\ K_3 &= [9.1360 \quad 5.6595 \quad 20.9195 \quad 1.3763] \end{aligned}$$



Then, The resolution of the optimization problem (34) permits the compute of the decentralized observation gain $L = \text{diag}(L_i), i = 1, 2, 3$, as follows:

$$L_1 = \begin{bmatrix} 2.3 \\ 1914.4 \\ 0 \\ -1.5 \end{bmatrix}, L_2 = \begin{bmatrix} 1.9 \\ 1368.1 \\ 0 \\ -1.2 \end{bmatrix}, L_3 = \begin{bmatrix} 1.9 \\ 1469.2 \\ 0 \\ -1.2 \end{bmatrix}$$

The effectiveness of the proposed control approach is demonstrated in the following simulations where two cases are studied:

- Case 1: we consider the autonomous augmented system, then the reference input vectors for the three reference models are null ($r_i = 0_{4 \times 1}$, $i = 1, 2, 3$).

In this case, the performances of the proposed H_∞ control scheme based on the decentralized observer, tested by numerical simulation, are shown in figures 1 to 4, on which are simulated the evolution of the real states and their reference model states : the rotor angle variation $\Delta\delta_i, i = 1, 2, 3$, the relative speeds w_1, w_2 and w_3 of the three generators, the mechanical power variation $\Delta P_{mi}, i = 1, 2, 3$, and the steam valve opening variation $\Delta X_{e1}, \Delta X_{e2}$ and ΔX_{e3} .

It appears on these curves that the decentralized observer based H_∞ model reference tracking control permits the stabilization of the real state variables and the improvement of the behavior of the interconnected system state variable in transient state.

- Case 2: the reference input vectors for the three reference models are given by:

$$r_i = \begin{bmatrix} 0 & 50\cos(0.8t) & 0 & 0 \end{bmatrix}^T, \quad i = 1, 2, 3$$

The reference input, having a cosinus trajectory, is applied on the relative speed to test the efficiency of the tracking control system with respect to a time varying reference which constitute a non easy tracking situation.

In this case, the performances of the proposed approach are shown in figures 5 to 8, on which are simulated the evolution of the real states and their reference model states: the rotor angle variation $\Delta\delta_i, i = 1, 2, 3$, the relative speeds $w_i, i = 1, 2, 3$, the mechanical power variation $\Delta P_{mi}, i = 1, 2, 3$, and the steam valve opening variation $\Delta X_{ei}, i = 1, 2, 3$, of the three generators.

It's clear from these curves that the decentralized observer based H_∞ model reference tracking control is efficient. It permits to the real state variables to reach the desired trajectories of the reference state variables of the three interconnected machine power system despite the strong disturbances emerging on the input reference models which are represented by a cosinus trajectory.

Figures 9 to 12 depict the behaviors of the observation errors of the four states: $\Delta\delta_i - \Delta\hat{\delta}_i$, $w_i - \hat{w}_i$, $\Delta P_{mi} - \Delta\hat{P}_{mi}$ and $\Delta X_{ei} - \Delta\hat{X}_{ei}$ for the three machines. It can be seen from these graphics that the

decentralized observer allows the reconstruction of the state variables of the multi-machine power system with a good performances. Indeed, the observed states converge rapidly towards the real state variables. In addition, the proposed observation approach permits to overcome the problem of sensitive and expensive sensors.

5 Conclusion

In this paper, a decentralized observer based H_∞ decentralized tracking control scheme has been proposed to ensure the stabilization of large scale interconnected systems. The H_∞ performances are also guaranteed despite the presence of the external disturbances and interconnections between subsystems, and the decentralized observer allows the reconstruction of the non measurable state variables of each subsystem. The proposed control approach leads to an optimization problem which has been formulated in terms of LMI constraints which resolution yields the control and the observation gains and the minimization of the tracking errors.

It has been shown from the simulation results that the proposed H_∞ decentralized tracking control scheme based on decentralized observer is efficient and permits the rapid stabilization of the global system, the reconstruction of the non measurable state variables and the guarantee of the desired performances despite strong perturbations applied to subsystems.

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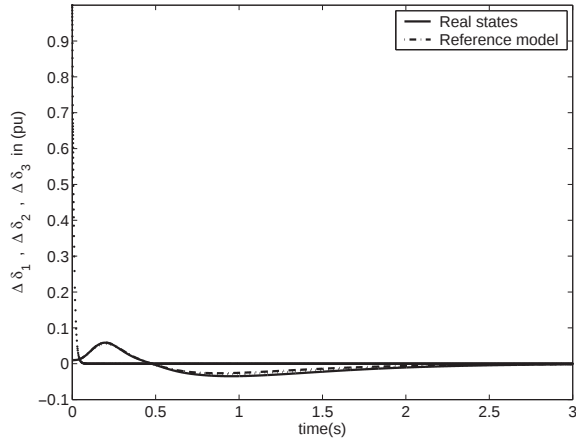


Figure 1: Behaviors of the real state variables $\Delta\delta_i$, $i = 1 \dots 3$, and their reference states of the three interconnected machines power system

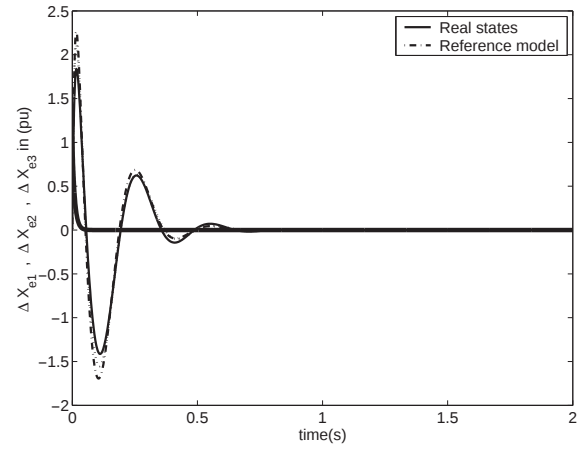


Figure 4: Behaviors of the real state variables ΔX_{ei} , $i = 1 \dots 3$, and their reference states of the three interconnected machines power system

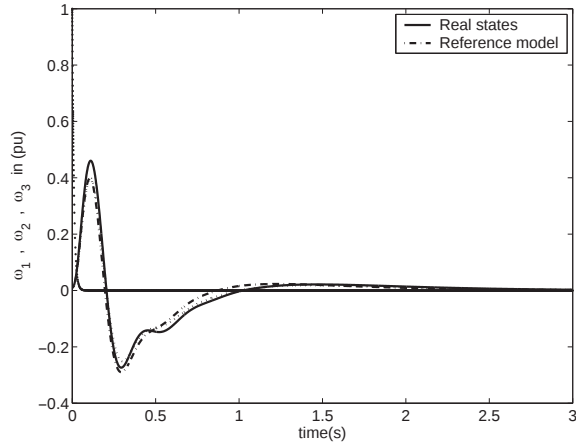


Figure 2: Behaviors of the real state variables ω_i , $i = 1 \dots 3$, and their reference states of the three interconnected machines power system

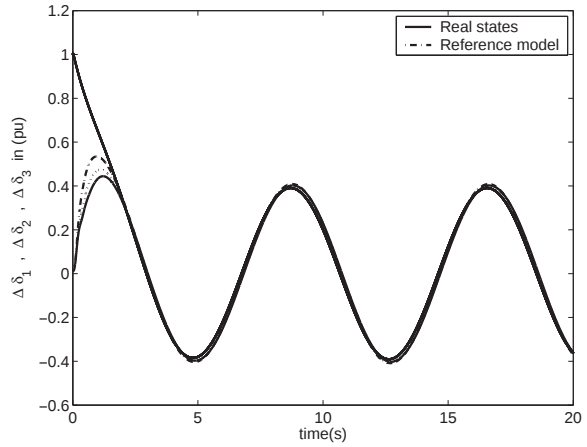


Figure 5: Behaviors of the real state variables $\Delta\delta_i$, $i = 1 \dots 3$, and their reference states of the three interconnected machines power system

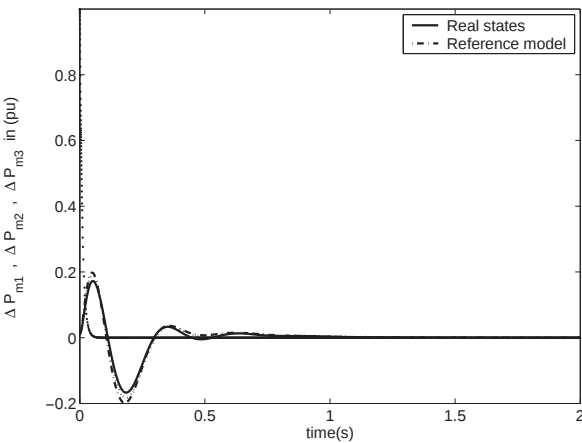


Figure 3: Behaviors of the real state variables ΔP_{mi} , $i = 1 \dots 3$, and their reference states of the three interconnected machines power system

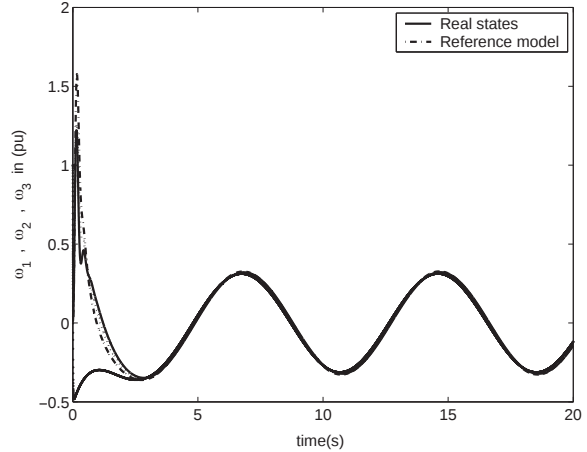


Figure 6: Behaviors of the real state variables ω_i , $i = 1 \dots 3$, and their reference states of the three interconnected machines power system



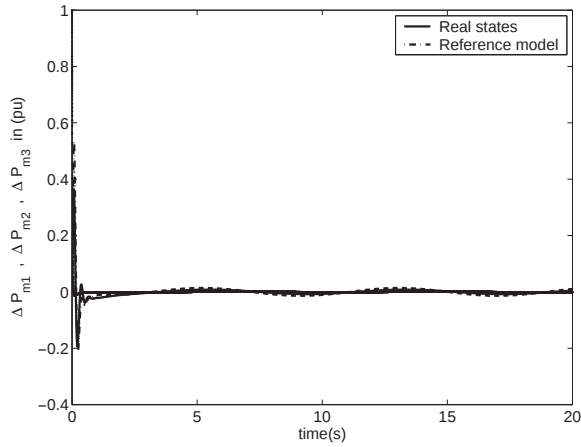


Figure 7: Behaviors of the real state variables $\Delta P_{mi}, i = 1 \dots 3$, and their reference states of the three interconnected machines power system

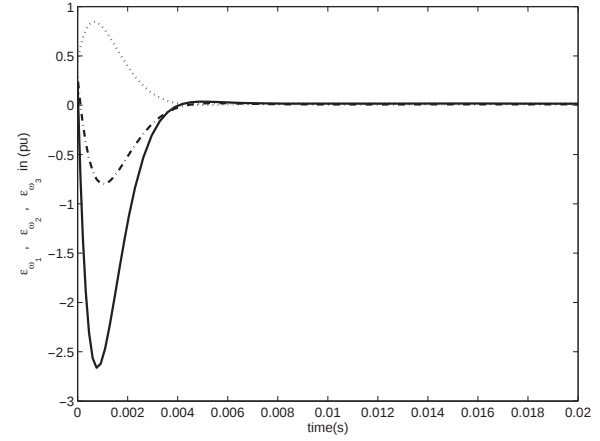


Figure 10: Behaviors of the observation errors of the relative speed of the three interconnected machines power system

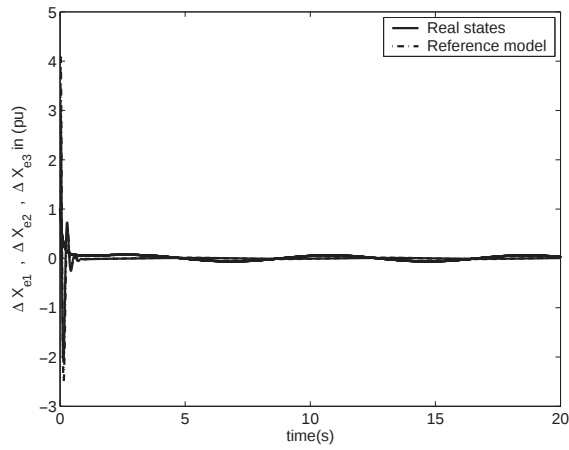


Figure 8: Behaviors of the real state variables $\Delta X_{ei}, i = 1 \dots 3$, and their reference states of the three interconnected machines power system

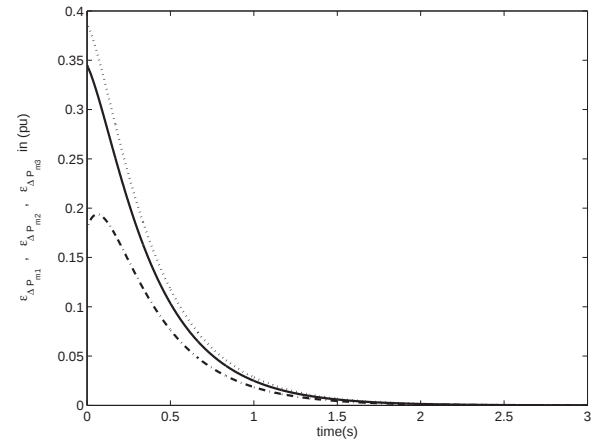


Figure 11: Behaviors of the observation errors of the mechanical power variation of the three interconnected machines power system

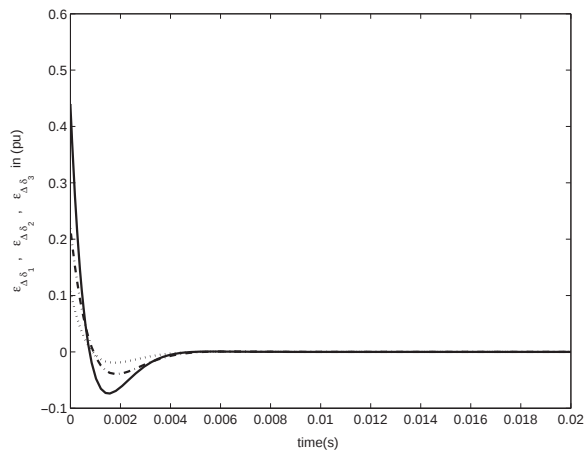


Figure 9: Behaviors of the observation errors of the rotor angle variation of the three interconnected machines power system

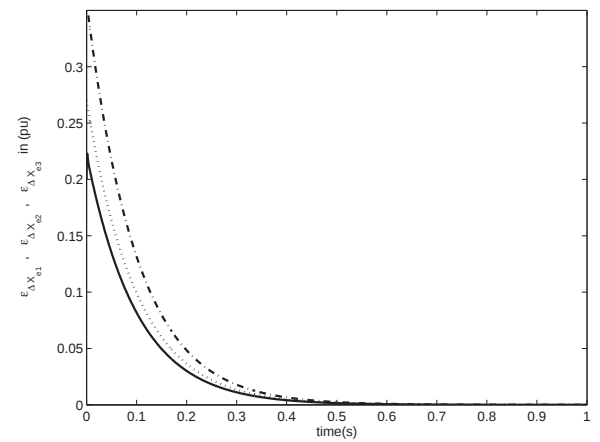


Figure 12: Behaviors of the observation errors of the steam valve opening of the three interconnected machines power system



6 Biographies

Sadok Dhbaibi

received the Master of Automatic from Ecole Supérieure des Sciences Techniques de Tunis in 2004. Currently, he is Assistant in Institut Supérieur des Etudes Technologiques de Mahdia and research member of Processes Study and Automatic Control Laboratory (LECAP) in the Ecole Polytechnique de Tunisie. Actually, he is preparing his PhD dissertation in Electrical Engineering in the field of decentralized observers based control of complex, nonlinear and uncertain systems with application on interconnected multi-machine power systems.



Université des Sciences et Techniques de Lille, France, in 1990, and the Doctorat d'Etat in Electrical Engineering from Ecole Nationale d'Ingénieurs de Tunis in 1995. Now, he is Professor of Electrical Engineering at the University of Tunis -Ecole Supérieure des Sciences et Techniques de Tunis-. He is also Director of the Processes Study and Automatic Control Laboratory (LECAP) at the Ecole Polytechnique de Tunisie. His domain of interest is related to the modelling, analysis and control of nonlinear systems with applications on electrical processes.

Ali Sghaier Tlili

received his Master of Systems Analysis and Numerical Treatment and his PhD in Electrical Engineering, both from Ecole Nationale d'Ingénieurs de Tunis in 1998 and 2003 respectively. He is currently an Assistant Professor in Faculté des Sciences de Bizerte and member of Processes Study and Automatic Control Laboratory (LECAP) in the Ecole Polytechnique de Tunisie. His current research interests include nonlinear, decentralized, multimodel and fuzzy observers based control for nonlinear, uncertain and complex systems with application on electromechanical processes.



Naceur Benhadj

Braiek was born in 1963. He obtained the Master of Electrical Engineering and the Master of Systems Analysis and Numerical Processing, both from Ecole Nationale d'Ingénieurs de Tunis in 1987, the Master of Automatic Control from Institut Industriel du Nord (Ecole Centrale de Lille) in 1988, the PhD in Automatic control from

